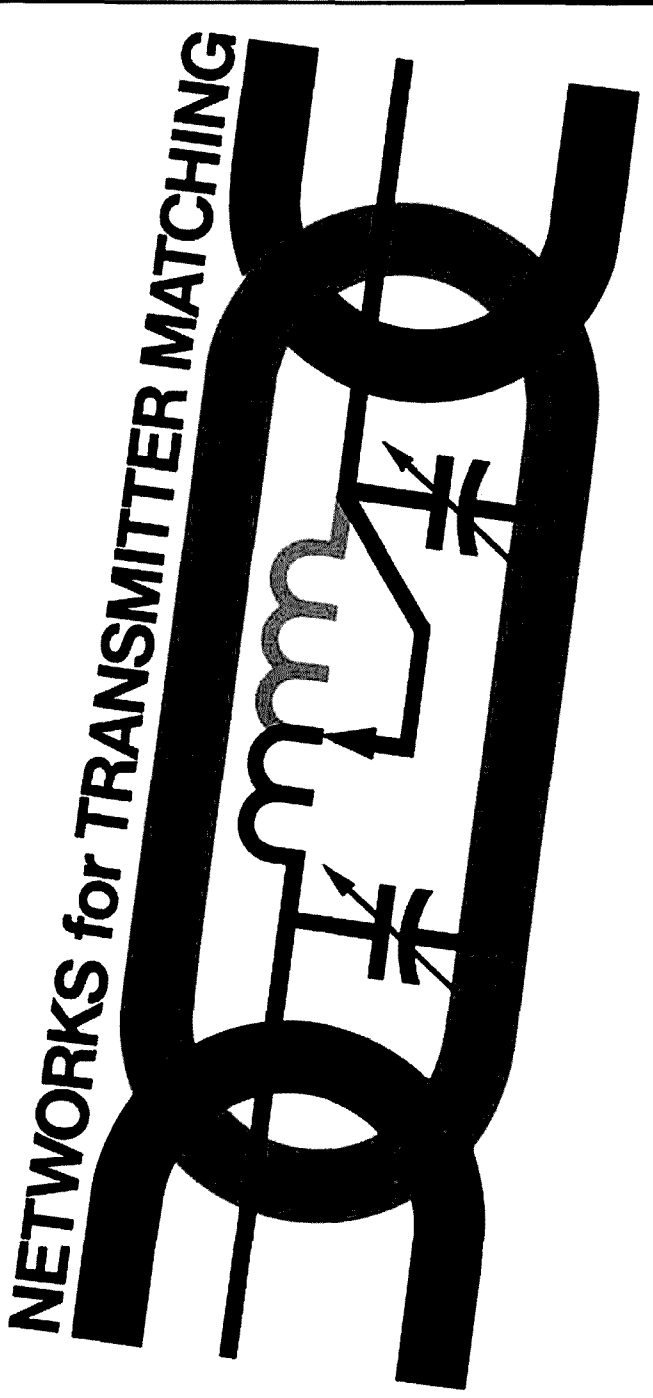
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ham radio magazine

JANUARY 1973



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- antenna tuning units 58

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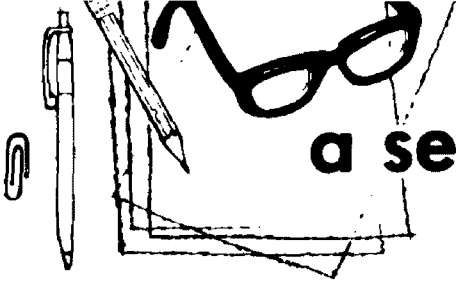
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a second look

by jim fisk

This is the time of the year when many high-school seniors are scurrying around, planning their future education, sending applications off to the college of their choice and taking entrance exams. Seniors who are also radio amateurs are probably considering a career in electronics. If they're lucky, they will have a knowledgeable guidance counselor who can steer them in the right direction; if not, they'll probably pick a school with a big name and work from there. Sometimes this works out and sometimes it doesn't — it depends entirely on what the student is looking for.

Electrical engineers who graduated before 1960 would probably not recognize the engineering curriculum now offered by their old alma mater because, during the past 15 years or so, engineering education has changed significantly. During this period (referred to by some as the post-Sputnik period), classical engineering education has tended to become less applied and more and more theoretically oriented. The backgrounds of some electrical engineering teaching staffs have changed from being primarily applied electronics to applied mathematics, and attempts to develop practical engineering programs have not been very successful.

In recent years the prestige of an engineering school has generally been gauged by the theoretical emphasis of its courses. Each school has tried to outdo the other in the theoretical sophistication of its curriculum. Unfortunately, the majority of the jobs in the sphere of electronics engineering do not require such an advanced mathematical sophistication as they do a "gut" understanding of electronics. If you talk to today's electrical engineering students, you will find that many of them do not know how to solve a simple steady-state ac problem, although they can invert a matrix and use state-variable techniques.

However, several colleges are now introducing four-year electronic technology programs in an attempt to get back to the old, practical engineering concept. The new Electronic Technology program at Trenton State College, for example, emphasizes electronics hardware and laboratory techniques as well as electrical theory. Unlike most engineering programs, at Trenton State the electrical courses begin in the first semester of the freshman year and continue right on through the senior year. And because of the applications orientation, graduates of two-year associate degree technician training programs can transfer into the B.S. program with little or no loss of credit.

The graduates of B.S. Technology programs have been pictured as fitting into the occupational spectrum somewhere between the technician and the engineer. However, many professors see technology graduates as having much wider employment opportunities. Actually, technology graduates should have many opportunities in areas of electrical applications and design traditionally occupied by engineering graduates, now vacated due to the change in emphasis of engineering education.

Students who are interested in this type of engineering education should be aware that there is a wide difference in B.S. programs offered under the title of "Technology." This condition is a result of the relative newness of the concept. Some programs are managerially oriented, others are applications and design oriented, while still others are little more than a two-year technician training program with added courses in the arts and humanities. Anyone desiring to enter this area should choose his school carefully to be sure he gets what he wants.

Jim Fisk, W1DTY
editor

networks for transmitter matching

Complete data on building networks for matching the impedance of your exciter to the impedance of your power amplifier

Irvin M. Hoff, W6FFC, 12130 Foothill Lane, Los Altos Hills, California 94022

Most transmitters are designed for 50-ohm output loads and the use of 50-ohm coax cable has become quite standard on most antenna systems used by amateurs. As the typical transmitter these days has 100 to 175 watts output, it is often used as an exciter to drive a linear amplifier to higher output power. These units normally are cathode-driven and are characterized by an input impedance that falls in the region of 20 to 200 ohms. Although in many cases the exciter can drive such an amplifier directly with satisfactory results, the use of a properly-terminated matching network can be most beneficial in a variety of ways: It allows maximum energy transfer (most output), presents the best load to the exciter, minimizes harmonic radiation (tvi, etc.) and allows barefoot operation without retuning.

Perhaps other advantages will come to mind. Some exciters have only a 50-ohm output, and cannot be retuned for other impedances.

input impedance

The input impedance of linear amplifiers is rarely the same from one band to another. Some amplifiers are not operated at zero-bias and actually drive the grid through a passive resistor. These systems, of course, usually present about the same impedance from one band to another, but are rarely 50 ohms to start with.

Formulas have been given to enable the calculation of the input impedance of a grounded-grid, cathode-driven amplifier. However, such formulas are all but worthless since they do not take the frequency into consideration. Measurements taken at the input of such amplifiers usually show a rather impressive variation from 10 to 80 meters, indicating that a formula would be quite misleading. These variations are caused by the manner in which the rf is isolated from the filament transformer (and hence the house wiring). Two methods are used to accomplish this: filament chokes, such as bifilar-wound coils, or low-capacitance filament transformers.

The best uniformity is normally obtained with the low-capacitance filament transformer, but such a transformer is not always available, and in any event would need to be mounted within a few inches of the tube base. This is not always convenient, so filament chokes are more commonly used. These chokes range from commercially-available units to homemade — the latter usually being two number-12 double-enameled wires wound simultaneously around a round ferrite rod until 11 turns (you would count 22 with the two wires) are on the rod. With proper bypassing these chokes allow the 60-Hz filament current to pass, but do not allow the high-frequency rf signal into the filament transformer.

Factors which seem to contribute to variations in input impedance from band to band include the voltage on the final amplifier, the type of tube or tubes being used, the frequency involved and the type of rf chokes used.

matching

I once had a Johnson Pacemaker 90-watt ssb transmitter. This unit could tune as high as 300 ohms on the output. I did not think any type of matching network to my linear was needed, but one day, while operating on 10 meters, I got a bad rf burn on my mouth when I came too close to the microphone. This led to an investigation of the input impedance, and I found on that particular transmitter it was only 15 ohms on 28 MHz; the Pacemaker could not handle this low impedance at all. A simple

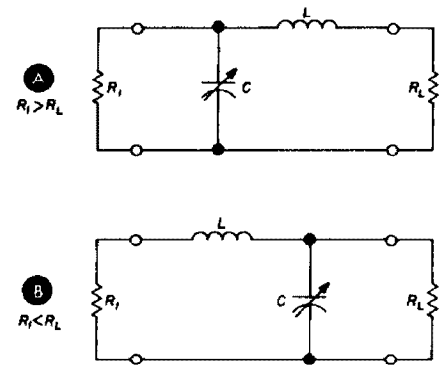


fig. 1. L-networks. The circuit in A is a step-down L-network; B shows a step-up L-network.

pi-network was used, and when incorporated for other bands, I found I not only had better output power, but could also then switch immediately from high power to barefoot, a distinct advantage over the previous system.

One company recommended that a particular length of coax should be used between the exciter and the amplifier. I personally always thought that this was a cop-out since it would be adequate (at best) on only one band!

Various articles have been written regarding the use of networks between the exciter and the linear, and this is now standard practice for most commercial units. These usually have input networks incorporated into the design, and are often adjustable if you wish to optimize them for your specific part of the band. They are usually switched automatically as you change the band selector.

Such networks are usually made up of pi-networks although a few use the more

simple L-network. The pi-newtork is usually preferred as greater control and uniformity are possible from band-to-band since the Q can be predetermined for consistent performance over a wider variety of impedances. The L-network is more simple, but at the same time it is somewhat more difficult to adjust for optimum swr.

networks

L-networks have been covered adequately in other texts, including the

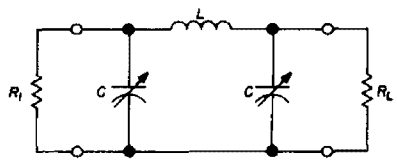


Fig. 2. A typical pi-network. R_i is the input load, R_L the output load.

ARRL Handbook, so only an example will be shown here (see fig. 1). Although this is a very simple circuit, it has several minor disadvantages.

For one thing, in the L-network Q cannot be controlled, and is usually very low. Also, if the network is used for all hf amateur bands, the capacitor often has to be switched from one end of the coil to the other. Further, the L-network has very little exciter loading due to the low Q and it offers very little harmonic suppression.

A typical pi-network is shown in fig. 2. It offers predictable performance as the Q may be preselected. It also offers additional harmonic suppression, presents a good load for exciter stability and can easily be used for all hf amateur bands.

input impedance

The input impedance of the network may be determined by testing; use of formulas should be avoided because the calculations rarely approximate the observed results.

The easiest and quickest method of measuring input impedance would be to use a variable impedance bridge, such as the long-since discountinued Heath AM-1. The ARRL Handbook also contains an excellent rf impedance bridge that may be easily built. These rf impedance bridges are basically a small swr bridge with a variable leg in the bridge so you can match the load impedance. Since an rf impedance bridge usually takes only a few milliwatts of power, they are easily driven from a grid-dip meter or ssb transmitter with the output cranked down.

Sufficient rf is introduced (with the load disconnected) to give either fullscale meter reading or nearly so. The load is then connected and the knob dialed for minimum meter reading. The impedance is then read directly from the calibrated dial. The high voltage must be running on the amplifier, and the meter hooked as close as possible to the place the network will be added.

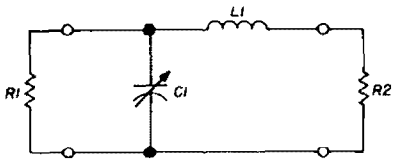
There are probably no typical impedances, but as a general rule I have found that most amplifiers I tested fell in to the neighborhood of 150 to 200 ohms on 80 meters, and around 15 to 30 ohms on 10 meters. The rest of the bands came somewhere in between. In many cases 20 meters offers a fairly decent match with no network at all.

If the input impedance is measured directly at the filament of the power tube it will be considerably less than 50 ohms on ten meters, and considerably more than 50 ohms on 80. The data shown below is for my own 4-1000A linear with

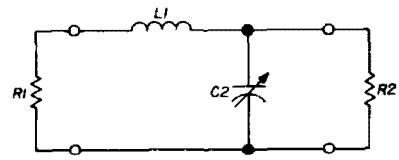
	impedance at tube base (ohms)	impedance at network (ohms)
80 meters	180	100
40 meters	155	60
20 meters	75	22
15 meters	50	40
10 meters	40	65

6000 volts on the plate. The amplifier uses a low-capacitance filament transformer. The first column of figures is the impedance measured right at the tube base; the second column shows the impedance at the end of a 6-foot piece of

table 1. L-network component values. Data is for matching a 50-ohm transmitter to a cathode-driven amplifier. The Q is set by the ratio of the input and output impedances and is shown for approximately the middle of each amateur radiotelephone band. The Q at the top of the band would be slightly less, at the bottom of the band it would be slightly greater.



R1 OHMS	F MHZ	C1 PF	L1 UH	C2 PF	R2 OHMS	"Q" QUAL.	R1 OHMS	F MHZ	C1 PF	L1 UH	C2 PF	R2 OHMS	"Q" QUAL.
50	1.9	4188	2.09	--	10	2.0	50	1.9	3420	5.13	--	30	0.8
50	3.8	2094	1.05	--	10	2.0	50	3.8	1710	2.56	--	30	0.8
50	7.2	1105	0.55	--	10	2.0	50	7.2	902	1.35	--	30	0.8
50	14.2	560	0.28	--	10	2.0	50	14.2	458	0.69	--	30	0.8
50	21.2	375	0.19	--	10	2.0	50	21.2	306	0.46	--	30	0.8
50	28.5	279	0.14	--	10	2.0	50	28.5	228	0.34	--	30	0.8
50	1.9	3420	3.42	--	20	1.2	50	1.9	4188	8.38	--	40	0.5
50	3.8	1710	1.71	--	20	1.2	50	3.8	2094	4.19	--	40	0.5
50	7.2	902	0.90	--	20	1.2	50	7.2	1105	2.21	--	40	0.5
50	14.2	458	0.46	--	20	1.2	50	14.2	560	1.12	--	40	0.5
50	21.2	306	0.31	--	20	1.2	50	21.2	375	0.75	--	40	0.5
50	28.5	228	0.23	--	20	1.2	50	28.5	279	0.56	--	40	0.5



50	1.9	--	11.24	3746	60	0.4	50	1.9	--	9.04	1129	160	1.5
50	3.8	--	5.62	1873	60	0.4	50	3.8	--	4.52	565	160	1.5
50	7.2	--	2.97	989	60	0.4	50	7.2	--	2.38	298	160	1.5
50	14.2	--	1.50	501	60	0.4	50	14.2	--	1.21	151	160	1.5
50	21.2	--	1.01	336	60	0.4	50	21.2	--	0.81	101	160	1.5
50	28.5	--	0.75	250	60	0.4	50	28.5	--	0.60	75	160	1.5
50	1.9	--	9.27	2649	70	0.6	50	1.9	--	9.19	1081	170	1.5
50	3.8	--	4.64	1324	70	0.6	50	3.8	--	4.60	541	170	1.5
50	7.2	--	2.45	699	70	0.6	50	7.2	--	2.43	285	170	1.5
50	14.2	--	1.24	354	70	0.6	50	14.2	--	1.23	145	170	1.5
50	21.2	--	0.83	237	70	0.6	50	21.2	--	0.82	97	170	1.5
50	28.5	--	0.62	177	70	0.6	50	28.5	--	0.61	72	170	1.5
50	1.9	--	8.65	2163	80	0.8	50	1.9	--	9.35	1039	180	1.6
50	3.8	--	4.33	1081	80	0.8	50	3.8	--	4.68	519	180	1.6
50	7.2	--	2.28	571	80	0.8	50	7.2	--	2.47	274	180	1.6
50	14.2	--	1.16	289	80	0.8	50	14.2	--	1.25	139	180	1.6
50	21.2	--	0.78	194	80	0.8	50	21.2	--	0.84	93	180	1.6
50	28.5	--	0.58	144	80	0.8	50	28.5	--	0.62	69	180	1.6
50	1.9	--	8.43	1873	90	0.9	50	1.9	--	9.51	1001	190	1.7
50	3.8	--	4.21	937	90	0.9	50	3.8	--	4.76	501	190	1.7
50	7.2	--	2.22	494	90	0.9	50	7.2	--	2.51	264	190	1.7
50	14.2	--	1.13	251	90	0.9	50	14.2	--	1.27	134	190	1.7
50	21.2	--	0.76	168	90	0.9	50	21.2	--	0.85	90	190	1.7
50	28.5	--	0.56	125	90	0.9	50	28.5	--	0.63	67	190	1.7
50	1.9	--	8.38	1675	100	1.0	50	1.9	--	9.67	967	200	1.7
50	3.8	--	4.19	838	100	1.0	50	3.8	--	4.84	484	200	1.7
50	7.2	--	2.21	442	100	1.0	50	7.2	--	2.55	255	200	1.7
50	14.2	--	1.12	224	100	1.0	50	14.2	--	1.29	129	200	1.7
50	21.2	--	0.75	150	100	1.0	50	21.2	--	0.87	87	200	1.7
50	28.5	--	0.56	112	100	1.0	50	28.5	--	0.64	64	200	1.7
50	1.9	--	8.41	1529	110	1.1	50	1.9	--	9.83	937	210	1.8
50	3.8	--	4.21	765	110	1.1	50	3.8	--	4.92	468	210	1.8
50	7.2	--	2.22	404	110	1.1	50	7.2	--	2.59	247	210	1.8
50	14.2	--	1.13	205	110	1.1	50	14.2	--	1.32	125	210	1.8
50	21.2	--	0.75	137	110	1.1	50	21.2	--	0.88	84	210	1.8
50	28.5	--	0.56	102	110	1.1	50	28.5	--	0.66	62	210	1.8
50	1.9	--	8.50	1416	120	1.2	50	1.9	--	9.99	909	220	1.8
50	3.8	--	4.25	708	120	1.2	50	3.8	--	5.00	454	220	1.8
50	7.2	--	2.24	374	120	1.2	50	7.2	--	2.64	240	220	1.8
50	14.2	--	1.14	189	120	1.2	50	14.2	--	1.34	122	220	1.8
50	21.2	--	0.76	127	120	1.2	50	21.2	--	0.90	81	220	1.8
50	28.5	--	0.57	94	120	1.2	50	28.5	--	0.67	61	220	1.8
50	1.9	--	8.61	1324	130	1.3	50	1.9	--	10.15	883	230	1.9
50	3.8	--	4.30	662	130	1.3	50	3.8	--	5.08	441	230	1.9
50	7.2	--	2.27	350	130	1.3	50	7.2	--	2.68	233	230	1.9
50	14.2	--	1.15	177	130	1.3	50	14.2	--	1.36	118	230	1.9
50	21.2	--	0.77	119	130	1.3	50	21.2	--	0.91	79	230	1.9
50	28.5	--	0.57	88	130	1.3	50	28.5	--	0.68	59	230	1.9
50	1.9	--	8.74	1249	140	1.3	50	1.9	--	10.31	859	240	1.9
50	3.8	--	4.37	624	140	1.3	50	3.8	--	5.16	430	240	1.9
50	7.2	--	2.31	330	140	1.3	50	7.2	--	2.72	227	240	1.9
50	14.2	--	1.17	167	140	1.3	50	14.2	--	1.38	115	240	1.9
50	21.2	--	0.78	112	140	1.3	50	21.2	--	0.92	77	240	1.9
50	28.5	--	0.58	83	140	1.3	50	28.5	--	0.69	57	240	1.9
50	1.9	--	8.88	1185	150	1.4	50	1.9	--	10.47	838	250	2.0
50	3.8	--	4.44	592	150	1.4	50	3.8	--	5.24	419	250	2.0
50	7.2	--	2.34	315	150	1.4	50	7.2	--	2.76	221	250	2.0
50	14.2	--	1.19	159	150	1.4	50	14.2	--	1.40	112	250	2.0
50	21.2	--	0.80	106	150	1.4	50	21.2	--	0.94	75	250	2.0
50	28.5	--	0.59	79	150	1.4	50	28.5	--	0.70	56	250	2.0

RG-58A/U where my matching network is placed.

You can instantly see the futility in trying to cut a piece of coax to just the right length to provide proper matching on a number of different bands. This table also illustrates how unpredictable it would be to try to use a formula to find the impedance!

In one rig I built, using a pair of 813s and a commercial FC-30 filament choke, the impedance varied widely, from 12 ohms on ten meters to over 200 ohms on 80 meters. Replacing the commercial filament choke with a homemade bifilar-wound unit gave results that varied much less, from about 30 ohms minimum on one band to 130 ohms on 80 meters. These figures are given only to illustrate the wide impedance variations possible from 3.5 through 29 MHz, and are unlikely to be typical of what you may experience with your own particular amplifier.

wattmeter method

The majority of you will not have access to an rf impedance bridge. You can still match the exciter to the amplifier, but it will take longer. The name of the game is low swr between the two units, so a wattmeter makes a good trial-and-error method of initially tuning the network. Once the settings have been found, you can mark them on the box and paste on tabs or use the sheet of paper I use.

In this case you observe, from the computer charts, the approximate inductance and capacitance, and start out by setting the inductance somewhere near what you think would be appropriate. With about half-power on the transmitter, rotate the variable capacitors while observing the reflected power. If it does not go to zero, tap up or down on the inductor and try again (the tap on the coil should be temporary until properly selected). This same technique is used on each different band.

using a swr bridge

This is the least desirable of the various methods. It will usually work, but is the most time-consuming of all and can be misleading. If you think you have gotten it just right, switch to the exciter barefoot and see if the antenna presents approximately the same load, plate current, output power, etc. without returning the exciter. This will provide a check on your accuracy, and is, of course, the desired end result anyway – the ability to switch from antenna to amplifier with similar results.

network placement

In commercial rf power amplifiers the matching network is usually quite near the tubes in the amplifier, and usually there is a separate network for each band. The appropriate network is switched in automatically with the band-selector knob.

It is not at all necessary to have the networks in the same cabinet with the rest of the transmitter. You may find it considerably more convenient to have it a few feet away from the amplifier where a simple network can be changed quickly whenever you bandswitch. This is the

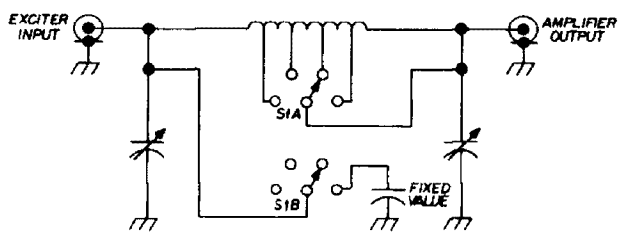
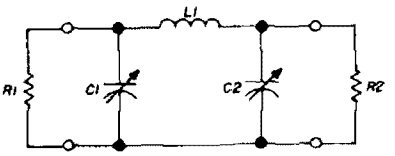


fig. 3 A typical pi-network for transmitter matching. The switch selects the proper tap for the various bands; the second switch section may be used to switch in parallel fixed values on the lower frequency bands.

arrangement I have used successfully for a number of years. I have a short piece of coax connecting the network minibox to the input of the amplifier. The length of the coax is in no way critical, but once the network is adjusted, of course, the coax length should then remain the same.

A piece of paper was temporarily placed on the front panel of the minibox, the correct settings for the various bands found and the paper marked. Then a

table 2. Pi-network component values. Data is for matching a 50-ohm transmitter to a cathode-driven amplifier. The Q has been chosen quite low to obtain broadband characteristics. The Q figure in the last column shows the worst-case condition at the bottom of the band using the inductance value shown.



R1 OHMS	F MHZ	C1 PF	L1 UH	C2 PF	R2 OHMS	"Q" QUAL.
50	1.8	4909	1.85	7612	10	3.6
50	3.5	2600	0.94	4153	10	3.8
50	7.0	1179	0.49	1678	10	3.3
50	14.0	579	0.25	801	10	3.2
50	21.0	384	0.16	528	10	3.2
50	28.0	300	0.12	439	10	3.4
50	1.8	3828	2.56	4992	20	3.3
50	3.5	2028	1.30	2679	20	3.4
50	7.0	920	0.68	1157	20	3.0
50	14.0	451	0.34	562	20	3.0
50	21.0	300	0.23	372	20	3.0
50	28.0	234	0.17	298	20	3.1
50	1.8	3403	3.08	3978	30	3.3
50	3.5	1802	1.56	2120	30	3.4
50	7.0	817	0.82	940	30	3.0
50	14.0	401	0.41	459	30	3.0
50	21.0	266	0.27	305	30	3.0
50	28.0	208	0.20	241	30	3.1
50	1.8	3090	3.54	3510	40	3.2
50	3.5	1636	1.79	1757	40	3.3
50	7.0	742	0.94	790	40	3.0
50	14.0	364	0.47	387	40	3.0
50	21.0	242	0.32	257	40	3.0
50	28.0	189	0.23	202	40	3.1
50	1.8	2872	3.95	2872	50	3.2
50	3.5	1521	2.00	1521	50	3.3
50	7.0	690	1.04	690	50	3.0
50	14.0	339	0.53	339	50	3.0
50	21.0	225	0.35	225	50	3.0
50	28.0	176	0.26	176	50	3.1
50	1.8	2689	4.35	2542	60	3.2
50	3.5	1427	2.20	1346	60	3.3
50	7.0	643	1.15	611	60	3.0
50	14.0	315	0.58	300	60	3.0
50	21.0	209	0.39	199	60	3.0
50	28.0	157	0.29	149	60	3.0
50	1.8	2559	4.68	2306	70	3.3
50	3.5	1361	2.37	1221	70	3.4
50	7.0	609	1.24	554	70	3.0
50	14.0	298	0.62	272	70	3.0
50	21.0	198	0.42	181	70	3.0
50	28.0	148	0.31	135	70	3.0
50	1.8	2432	5.01	2104	80	3.3
50	3.5	1295	2.54	1114	80	3.4
50	7.0	576	1.33	506	80	3.0
50	14.0	281	0.67	248	80	3.0
50	21.0	187	0.45	165	80	3.0
50	28.0	140	0.34	124	80	3.0
50	1.8	2319	5.34	1939	90	3.3
50	3.5	1237	2.71	1027	90	3.4
50	7.0	546	1.42	466	90	3.0
50	14.0	267	0.71	229	90	3.0
50	21.0	177	0.48	152	90	3.0
50	28.0	133	0.36	114	90	3.0
50	1.8	2216	5.66	1799	100	3.3
50	3.5	1184	2.87	953	100	3.4
50	7.0	520	1.50	432	100	3.0
50	14.0	253	0.76	212	100	3.0
50	21.0	168	0.50	141	100	3.0
50	28.0	126	0.38	106	100	3.0
50	1.8	2120	5.96	1679	110	3.3
50	3.5	1135	3.02	889	110	3.4
50	7.0	495	1.58	403	110	3.0
50	14.0	241	0.80	198	110	3.0
50	21.0	160	0.53	131	110	3.0
50	28.0	120	0.40	99	110	3.0
50	1.8	2036	6.26	1577	120	3.3
50	3.5	1092	3.17	835	120	3.4
50	7.0	473	1.66	379	120	3.0
50	14.0	230	0.84	186	120	3.0
50	21.0	152	0.56	123	120	3.0
50	28.0	114	0.42	93	120	3.0
50	1.8	1959	6.54	1488	130	3.3
50	3.5	1052	3.32	708	130	3.4
50	7.0	453	1.74	358	130	3.0
50	14.0	220	0.87	175	130	3.0
50	21.0	146	0.58	117	130	3.0
50	28.0	109	0.44	87	130	3.0
50	1.8	1887	6.82	1410	140	3.3
50	3.5	1015	3.46	747	140	3.4
50	7.0	434	1.81	339	140	3.0
50	14.0	210	0.91	166	140	3.0
50	21.0	139	0.61	110	140	3.0
50	28.0	104	0.46	83	140	3.0
50	1.8	1821	7.09	1342	150	3.3
50	3.5	981	3.60	710	150	3.4
50	7.0	417	1.88	322	150	3.0
50	14.0	201	0.95	158	150	3.0
50	21.0	133	0.63	105	150	3.0
50	28.0	100	0.47	79	150	3.0
50	1.8	1757	7.36	1279	160	3.3
50	3.5	949	3.73	678	160	3.4
50	7.0	400	1.95	307	160	3.0
50	14.0	193	0.98	151	160	3.0
50	21.0	128	0.66	100	160	3.0
50	28.0	96	0.49	75	160	3.0
50	1.8	1697	7.62	1223	170	3.3
50	3.5	918	3.86	648	170	3.4
50	7.0	384	2.02	294	170	3.0
50	14.0	185	1.02	144	170	3.0
50	21.0	122	0.68	96	170	3.0
50	28.0	92	0.51	72	170	3.0
50	1.8	1641	7.88	1173	180	3.3
50	3.5	890	3.99	621	180	3.4
50	7.0	369	2.08	282	180	3.0
50	14.0	178	1.05	138	180	3.0
50	21.0	117	0.70	92	180	3.0
50	28.0	88	0.53	69	180	3.0
50	1.8	1589	8.12	1127	190	3.3
50	3.5	865	4.12	597	190	3.4
50	7.0	355	2.15	271	190	3.0
50	14.0	170	1.08	133	190	3.0
50	21.0	113	0.72	88	190	3.0
50	28.0	84	0.54	66	190	3.0
50	1.8	1538	8.37	1085	200	3.3
50	3.5	837	4.24	575	200	3.4
50	7.0	342	2.21	261	200	3.0
50	14.0	163	1.11	128	200	3.0
50	21.0	108	0.74	85	200	3.0
50	28.0	81	0.56	64	200	3.0
50	1.8	1512	8.58	1054	210	3.4
50	3.5	824	4.35	558	210	3.5
50	7.0	335	2.27	253	210	3.1
50	14.0	160	1.14	124	210	3.0
50	21.0	105	0.76	83	210	3.0
50	28.0	79	0.57	62	210	3.0
50	1.8	1515	8.75	1035	220	3.4
50	3.5	826	4.43	548	220	3.6
50	7.0	335	2.32	249	220	3.1
50	14.0	160	1.17	122	220	3.1
50	21.0	106	0.78	81	220	3.0
50	28.0	79	0.58	61	220	3.0
50	1.8	1518	8.91	1016	230	3.5
50	3.5	828	4.52	538	230	3.6
50	7.0	335	2.36	244	230	3.2
50	14.0	160	1.19	120	230	3.1
50	21.0	106	0.79	80	230	3.1
50	28.0	79	0.60	60	230	3.1
50	1.8	1520	9.08	999	240	3.6
50	3.5	829	4.60	529	240	3.7
50	7.0	335	2.41	240	240	3.3
50	14.0	160	1.21	118	240	3.2
50	21.0	106	0.81	78	240	3.2
50	28.0	79	0.61	59	240	3.2

nicer looking paper was drawn up with markings for those settings, typewritten with the band-markings, and attached to the minibox. This allows very rapid setting of the box whenever I bandswitch, yet only one coil and two variable capacitors are used.

Other methods may come to mind that will work adequately for your purpose. Trying to put the networks into the amplifier usually makes additional problems with regard to space, synchronizing with the bandswitch, etc. Thus, the remote minibox idea may appeal to some of you who do not have space in the amplifier or the technical capability of providing mechanical selection when the bandswitch is rotated.

components

Even with 100 watts output, there is only about 1.4 rf amps flowing. Consequently, rather small inductors, such as B&W stock can be used successfully. B&W type 3018 comes in 4-inch lengths, 8 turns per inch; the full 4 inches is 9.4 microhenries. Price is well under $2. B&W type 3014 is also 8 turns per inch, 3-inches long, and 4.8 microhenries; cost is approximately $1.50. These should give you ideas, and a wide variety of similar inductances are available.

Even with 100 watts output, the voltage across 50 ohms is only about 70 rms. Almost any type of variable capacitor, including the common 365 pF broadcast type will be more than adequate. You can easily find these for free from junker a-m radios of another era, and usually in gangs of two or three on the same shaft.

You will probably want a bandswitch for the network. Any type of switch capable of handling small amounts of rf will be adequate, and the additional pole/poles may be used to switch in fixed values for the lower frequencies, if desired. Ceramic or steatite switches are recommended.

Fixed capacitors should be rated for at least 150 or 200 volts, and capable of handling rf currents. Mica transmitting types are excellent. Low-cost door-knob capacitors are also good and are usually capable of handling kilowatt outputs.

Some commercial amplifiers use fixed capacitors and a slug-tuned variable inductor. Unless you have some means of determining the actual impedance to be matched, tuneup could be very time consuming, and fairly costly unless a large supply of capacitors suitable for rf is available. Also, many of the available slug-tuned inductors will not handle the amp or two of rf current without damage.

summary

Some method of matching the 50-ohm output impedance to the input of a linear amplifier should be offered. A good, simple but effective method is to build a single, variable pi-network and place it in a convenient place a few feet from the amplifier. A rf wattmeter may be used for initial tuneup, and simple markings placed on the box containing the network so rapid band changes can be made. Tables are included for both pi-networks and L-networks. These were computer-derived and include values for 1.9 through 29.7 MHz.

ham radio



"You're not interested in ham radio! How did that stupid computer ever match us up?"

digital readout variable-frequency oscillator

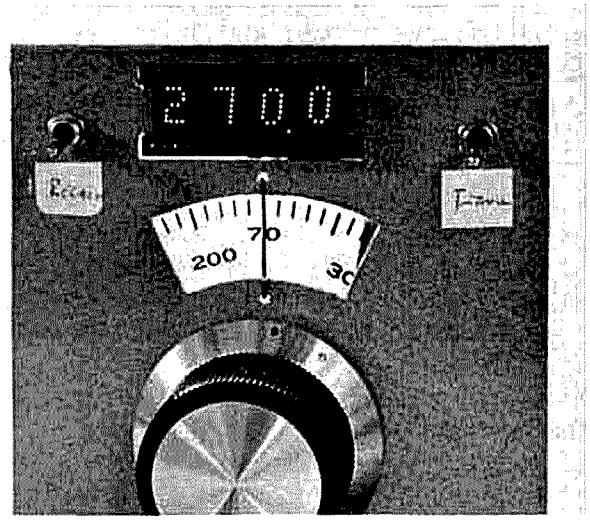
This digital readout vfo covers the range from 5.0 to 5.5 MHz and costs less than $100 to build

Gerd H. Schrick, WB8IFM, 4741 Harlou Drive, Dayton, Ohio 45432

Frequency counters have been around for some time, but only recently has the market become abundant with suitable parts and components making it possible to build a digital frequency station readout in a few evenings for under $100.00. The counter described here measures the frequency of the transceiver vfo or any other piece of equipment. It is designed for the Heath-kit HW100 which has a 5 to 5.5 MHz variable oscillator and requires a downcount for proper frequency reading. However, the design can easily be changed to accomodate other equipment.

The prototype was built into an external vfo which in turn is connected to my HW100. However, since it is very compact it can be incorporated into almost any existing equipment. The digital dial reads the frequency of the oscillator, indicating hundreds, tens, ones and a decimal of kilohertz. You have to remember the band, which in many cases is required anyway, for dial reading. A switch makes it possible to start reading at either 000.0 or at 500.0, taking care of the 75- and 10-meter bands which start out at 500.0. Another switch makes it possible to correct for errors that are introduced by inaccurate crystals.

The entire project (without power supply) consists of only 13 integrated circuits, four numerical readout ICs, two surplus crystals and a few resistors and



Front dial of the digital readout vfo.

capacitors. The following paragraphs will describe the various functions of the counter. The actual hardware consists of only three parts: a 3.75 x 6-inch Vero-board with the ICs and crystals, the 1 x 2-inch readout panel, which can be

cycle. A complete schematic of the time-base circuit and the times involved can be seen in fig. 4.

mixer

The 5- to 5.5-MHz output of the vfo is

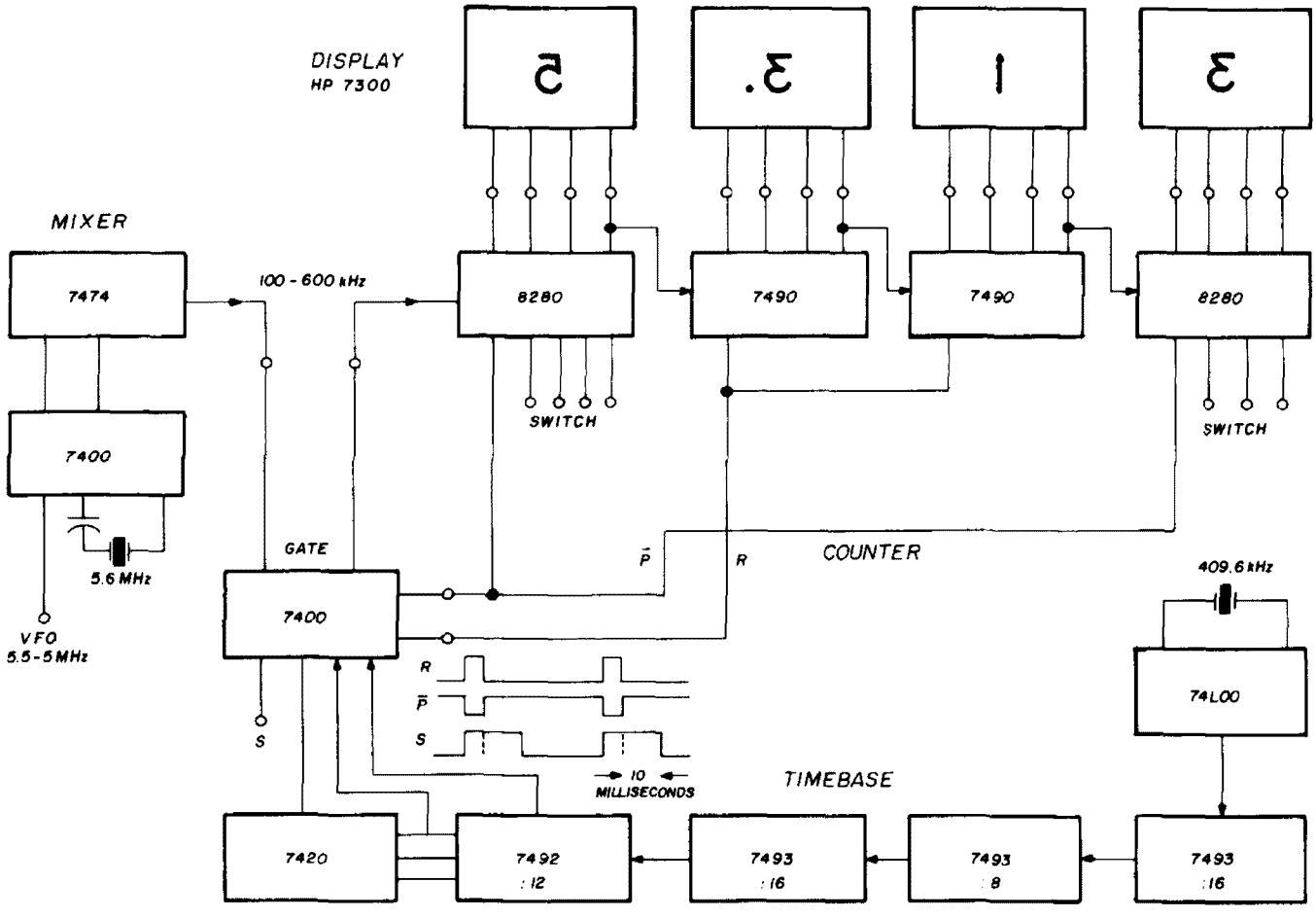


fig. 1. Block diagram of the digital readout vfo. All interconnections are shown except supply leads, grounds and switches.

mounted somewhere on the front panel of the equipment, and the power supply. The power supply described here uses an IC for regulation; however, any power supply of similar design can be used.

The oscillator uses a 74L00 IC, followed by three 7493 ICs and one 7492 IC connected as dividers (see fig. 1). The 7492 divide-by-12 IC plus two multiple-gate ICs (7420 and 7400) are used to derive the time-base functions: $\bar{P}$ (present), R (reset) and S (suppress). The bar over the P indicates a negative-going pulse. The suppress function turns the readout numbers off briefly and permits counting during the last two-thirds of the

not measured directly, but is mixed with another crystal oscillator. This not only reduces the accuracy requirements for the timebase; the entire counting process is relaxed since the actual frequency being measured is now only in the hundreds of kHz. Also, it is possible to choose the right crystal for the particular vfo and up or down counting scheme without changing anything else in the counter.

A rather unknown type of mixer using a D-type flip-flop (7474) is used (fig. 2). With this mixer there is no need for any external filter components whatsoever; moreover, there is no loss involved and consequently no need for amplification.

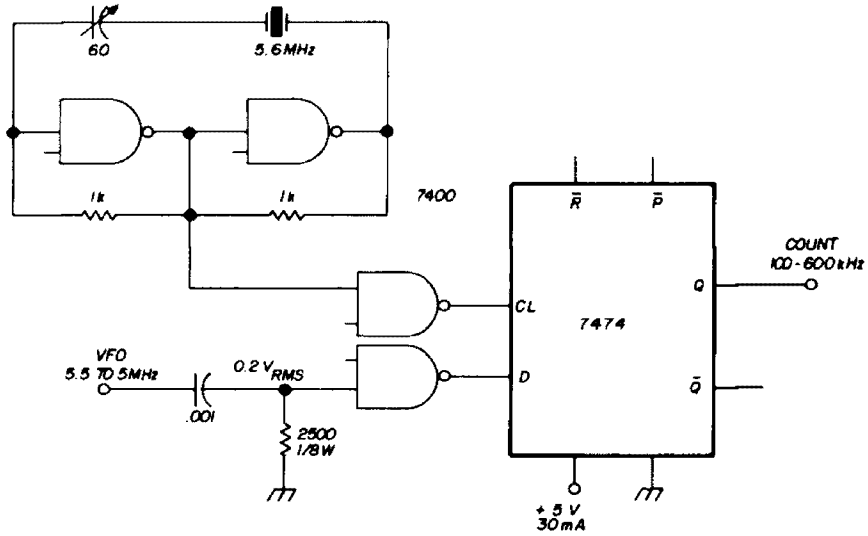


fig. 2. Digital mixer uses D-type flip-flop.

The oscillator uses a surplus 5600-kHz crystal and a 7400 IC to beat the 5- to 5.5-MHz vfo down to 100 to 600 kHz. The crystal oscillator can be adjusted with a trimmer capacitor to within ±300 Hz, so it can take care of some of the various band crystal frequency offsets.

counter

The counter is a simple ripple counter consisting of four individual decade counter ICs, fig. 3. The first and last ICs are 8280s or 74196s, and the two middle ones are 7490s. Each counter has a four-line (ABCD) output leading directly to the respective numeric indicators.

The first counter, which provides the last digit or hundreds of Hz, can be preset by properly grounding the four data lines D_A, D_B, D_C and D_D to any number from 0 to 9, thus permitting a correction (by separate switch) for individual bands. The last counter can be preset by switch to read from either 000.0 to 500.0 or from 500.0 to 1000.0. The middle counters have only normal reset capability. The individual counting process takes 10 milliseconds (1/100 of a second) and is repeated about 17 times per second, causing a slight flicker like a home movie.

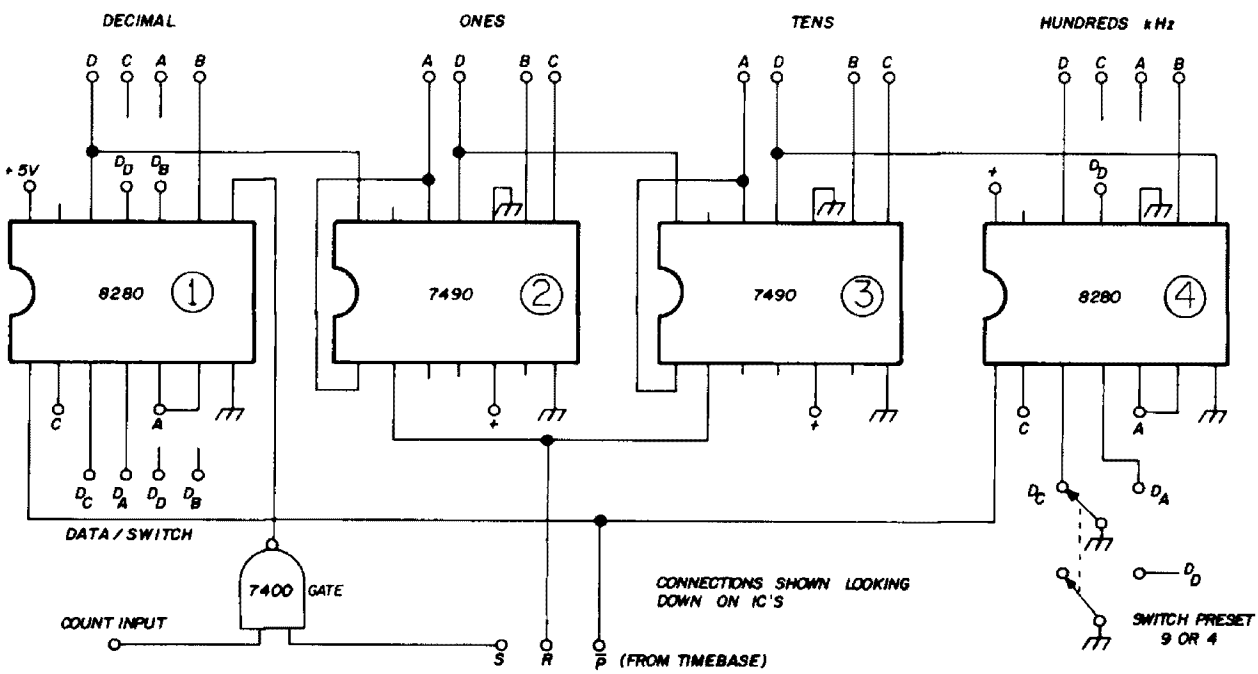


fig. 3. Four-stage ripple counter.

However, this is not at all annoying and indicates that the counter is operating properly. If, for instance, the flicker is not visible, the time base might be oscillating at a harmonic, thus giving an erroneous reading.

correction switch

The switch for correcting frequency offsets of inaccurate band crystals is a five-position, four-contact switch. The contacts correspond to the D_A, D_B, D_C and D_D data lines leading to the first counter chip, and the five positions correspond to the 75- through 10-meter amateur bands. Some people might want to use more positions to cover other 10-meter segments. The contact leads are brought out to a long socket capable of accepting double-sided circuit boards. Short pieces of wire are inserted between contacts according to the offset to be corrected and the BCD-code depicted in fig. 5.

power supply

This is a straight-forward power supply design. However, many similar schemes can be used. You might want to extract the power from 6.3-V filament winding. Approximately 600 mA at 5 V is required. An IC is used for regulation. This IC now costs about $2.00, looks and mounts like a T03 transistor and regulates any supply having between 7 and 25 V input to 5 volts output. Make sure you use several thousand microfarads for the filtering capacitor. The 2N3719 PNP-transistor (or similar) is used for switching the display. A separate ze-

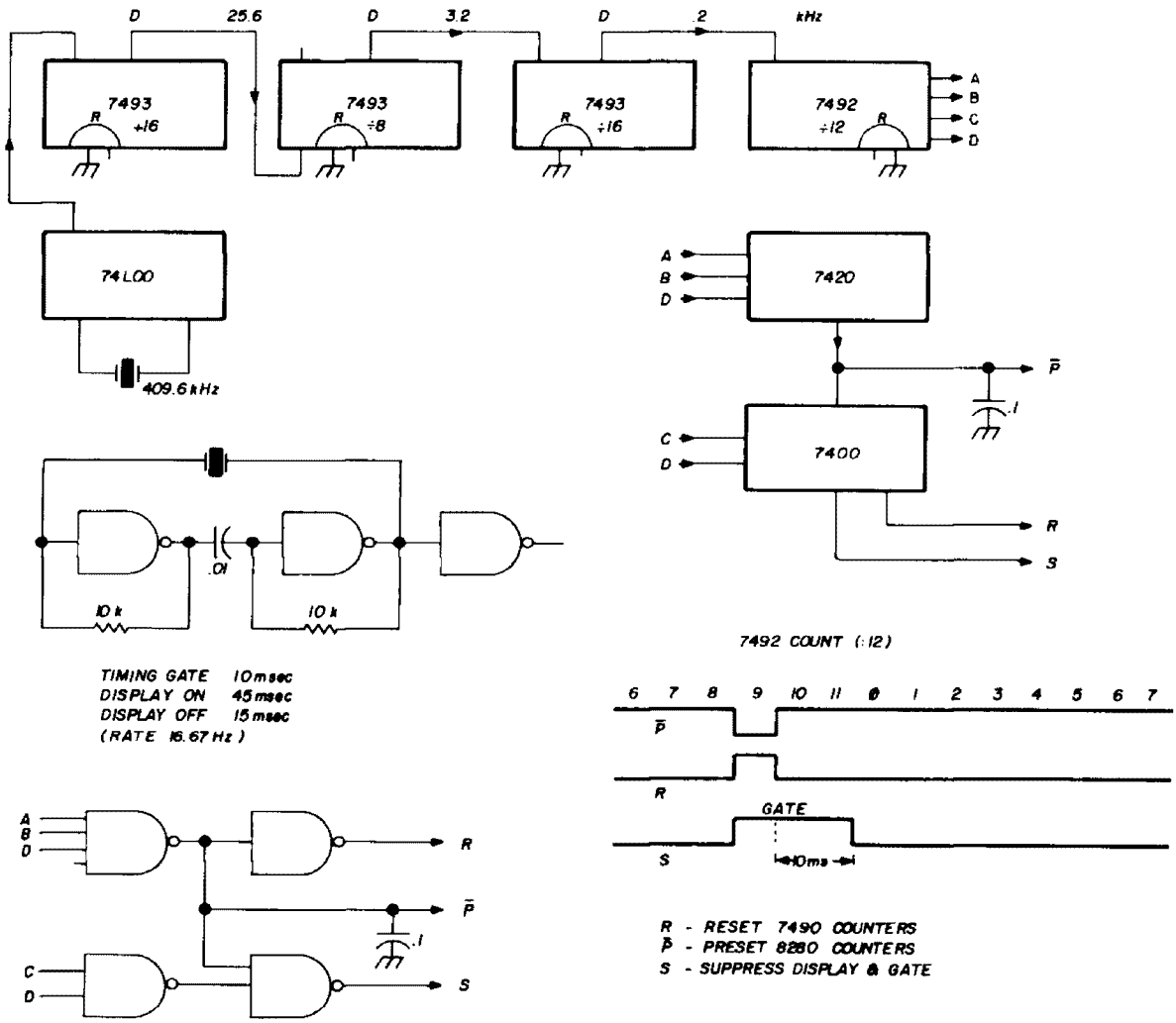


fig. 4. Complete timebase circuit for the digital readout vfo.

ner-regulated voltage is available for an external transistor vfo.

Hewlett-Packard type 5083-7300 numerical indicators were used in the unit.* These display numbers by utilizing a 4 x 7 light-emitting diode dot matrix. They accept a four-line BCD code from the actual counter. They also have a memory capability which is not being used in this counter.

Although each digit is only 0.2 by 0.3 inch, they are rather bright and can be seen clearly from up to 10 feet away. A missing dot, if it should occur, does not give an erroneous reading as would a missing section of a seven-segment readout. No extra power supply is required, nor is a decoder, as would be needed for almost all other designs.

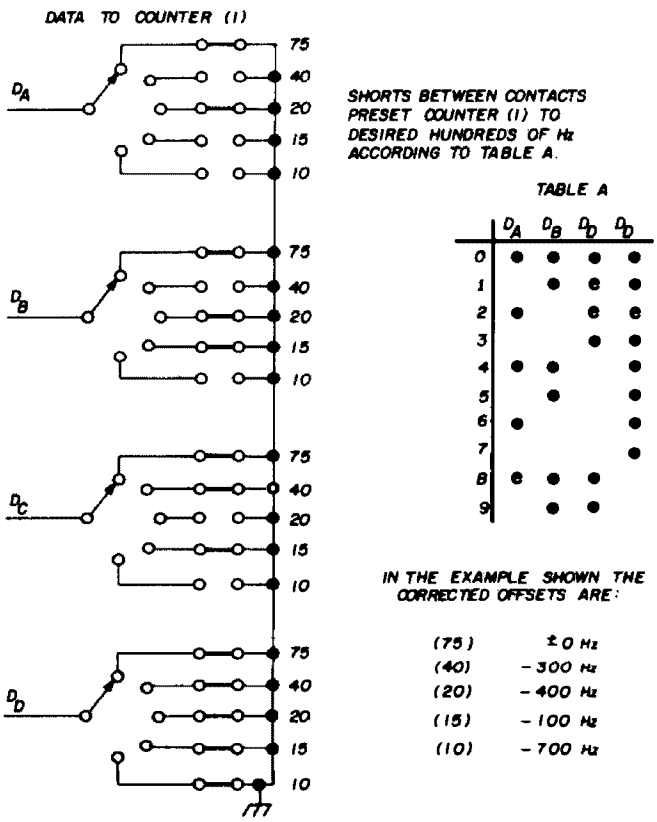


fig. 5. Bandswitch for correcting slight crystal frequency offsets.

*The Hewlett-Packard 7300 light-emitting digital display ICs are available from any H-P Sales Office. To find the address of your local Sales Office, look in the Yellow Pages, or write to Hewlett-Packard, 1501 Page Mill Road, Palo Alto, California 94304.

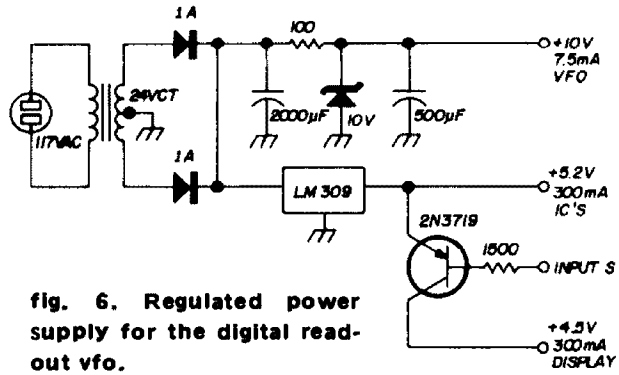


fig. 6. Regulated power supply for the digital readout vfo.

The Hewlett-Packard light-emitting readouts are somewhat more expensive than Nixie tubes, but they offer enough advantages to be well worth the price difference (the cost for one 7300 is $14.50). The readout can be mounted with a 36-pin IC socket, resulting in a rather flat, 1 x 2-inch display.

timebase

A frequency counter has to derive a gate from some standard, which is pretty much the same as using a crystal calibrator, except that the final result with the counter is a readout every 100 Hz and you do not have to search and zero beat those frequency marks.

This particular timebase starts off with a surplus crystal at 409.6 kHz and then, using a divide-by-two scheme, arrives at a gate which is 10 milliseconds wide (fig. 4). The number of Hz measured with this gate gives frequency directly with the last digit being in hundreds of Hertz. Other crystals and dividers can be used, providing you build an accurate 10-millisecond gate.

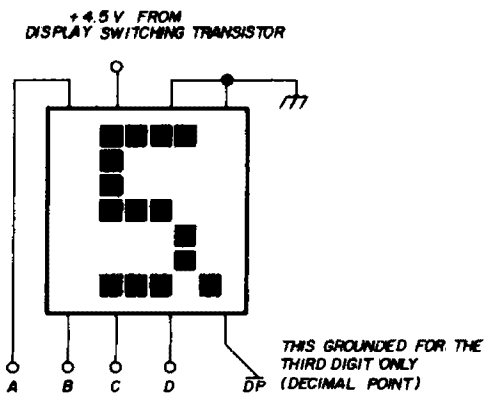


fig. 7. Wiring of light-emitting displays.

Although it is quite helpful to understand digital ICs, it is not absolutely necessary for this project. No special wiring diagram or printed circuit layout has been attempted. However, from the figures and schematics accompanying this article it should be relatively easy to perform the wiring chore; it can be done in two evenings.

It is recommended that you make major interconnections with IC socket pins and pieces of wire. Multicolor, no. 24 telephone wire is best suited; it should not be difficult to obtain scrap pieces.

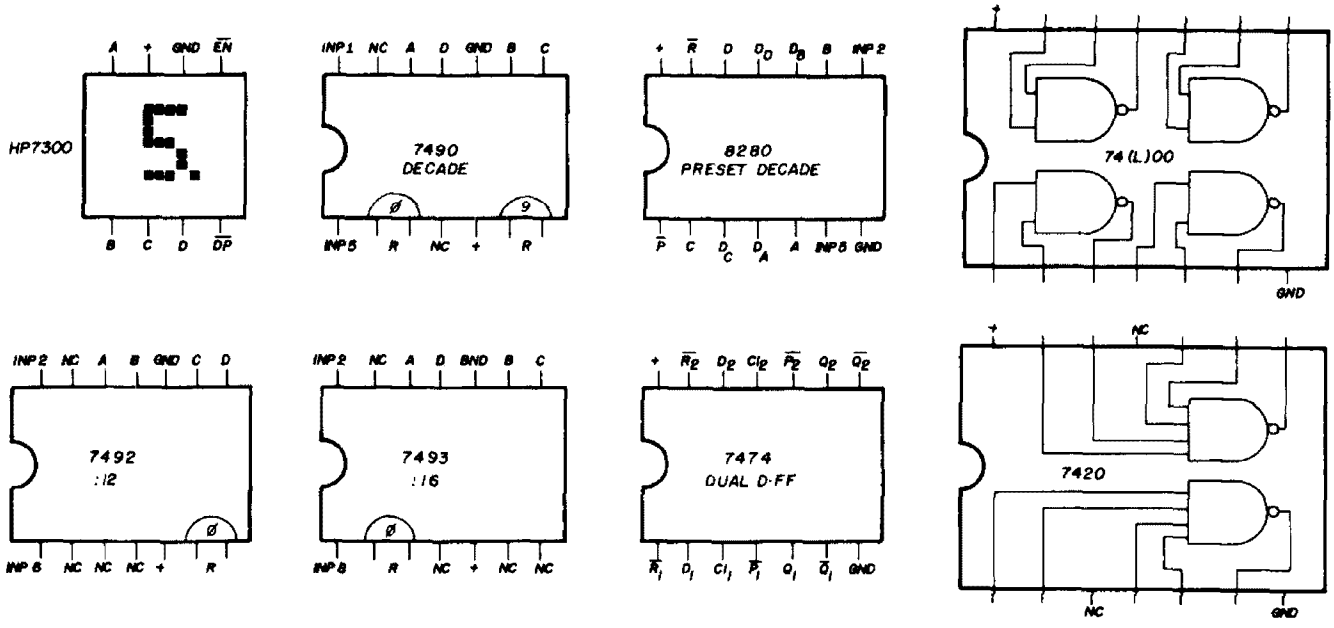


fig. 8. Base connections of the ICs used in the digital readout vfo (top views).

Great care must be exercised to ensure correct connections; every wire is important, components are rather small and IC lugs are spaced only 1/10 inch.

calibration

In the question of calibration a detailed procedure is being worked out. There are problems with the accuracy of the customary 100-kHz crystal and problems in detecting a beatnote on an ssb receiver that attenuates frequencies below 300 Hz. Also, there is the question of what exactly is the frequency of an ssb signal; is it, as the FCC questionaire implies, the carrier frequency, which of course is never transmitted, or is it a nominal center frequency in the middle of the sideband as used in Heathkit equipment?

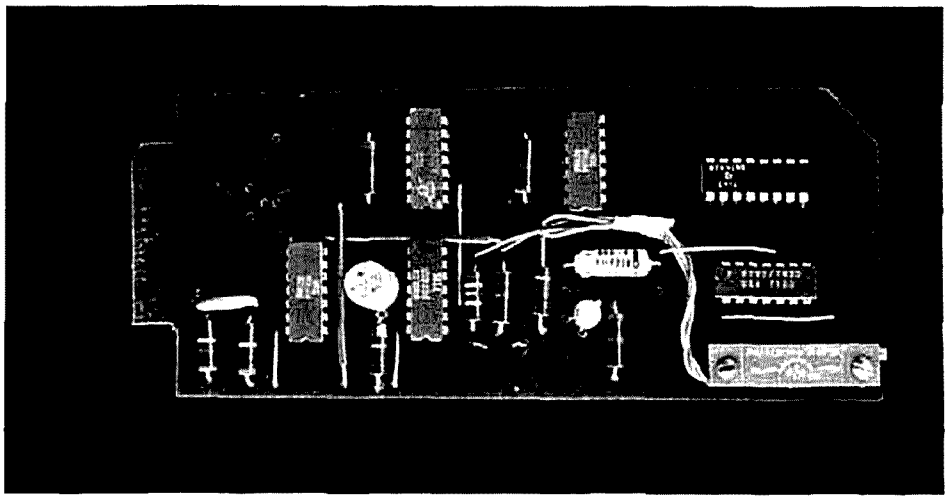
conclusion

The counter was first shown at the Dayton Hamvention in April, 1972, and has been in use ever since. You very soon become accustomed to reading the numbers, and although my vfo used the excellent Drake SPR4 dial, I hardly look at it any more. Quick departure from the operating frequency is possible. To look for a clear frequency, for example, and in an instant you can be back within a 100 Hertz of the old frequency, almost making the external vfo scheme or the clarifier obsolete.

One final comment about the improvement in frequency reading. First: Once the counter is checked out and calibrated there is no need for recalibration since only crystals are used for reading the frequency. Second: Compared to the ±1 kHz of the Drake dial the accuracy is now ±0.1 kHz, a ten-fold improvement over the very good Drake dial.

This counter exhibits the usual ±1 ambiguity in the last digit. However, this can be avoided by synchronizing the gate with the incoming frequency. A conversion taking care of this problem will be forthcoming.

ham radio



automatic line feed for RTTY

This low-cost, all solid-state automatic line-feed prevents end-of-line print garble and overprinting

Bert Kelley, K4EEU, 2307 South Clark Avenue, Tampa, Florida

When an RTTY transmission begins, there is no assurance that the receiving printer will start on the same line as the transmitting printer. Sometimes the line feed character is garbled, occasionally it's not even sent — these are all causes for loss of copy (or a "black box" of overstrikes on the right margin).

When no line feed is received, the printer runs up to the right margin and stops there, overprinting characters until a valid line-feed signal is received. This copy may be missed even if the operator is present, because attempts to manually inject a line-feed signal through the local loop may fail due to incoming signal interference.

An automatic RTTY line-feed system greatly improves the quality of the received copy, and is a definite operating convenience. There are various mechanical line-feed systems available, but they are unreliable and sometimes difficult to adjust.[1,2] The all solid-state system described here has proved to be very reliable, requires only one simple adjustment, and doesn't cost an arm and a leg.

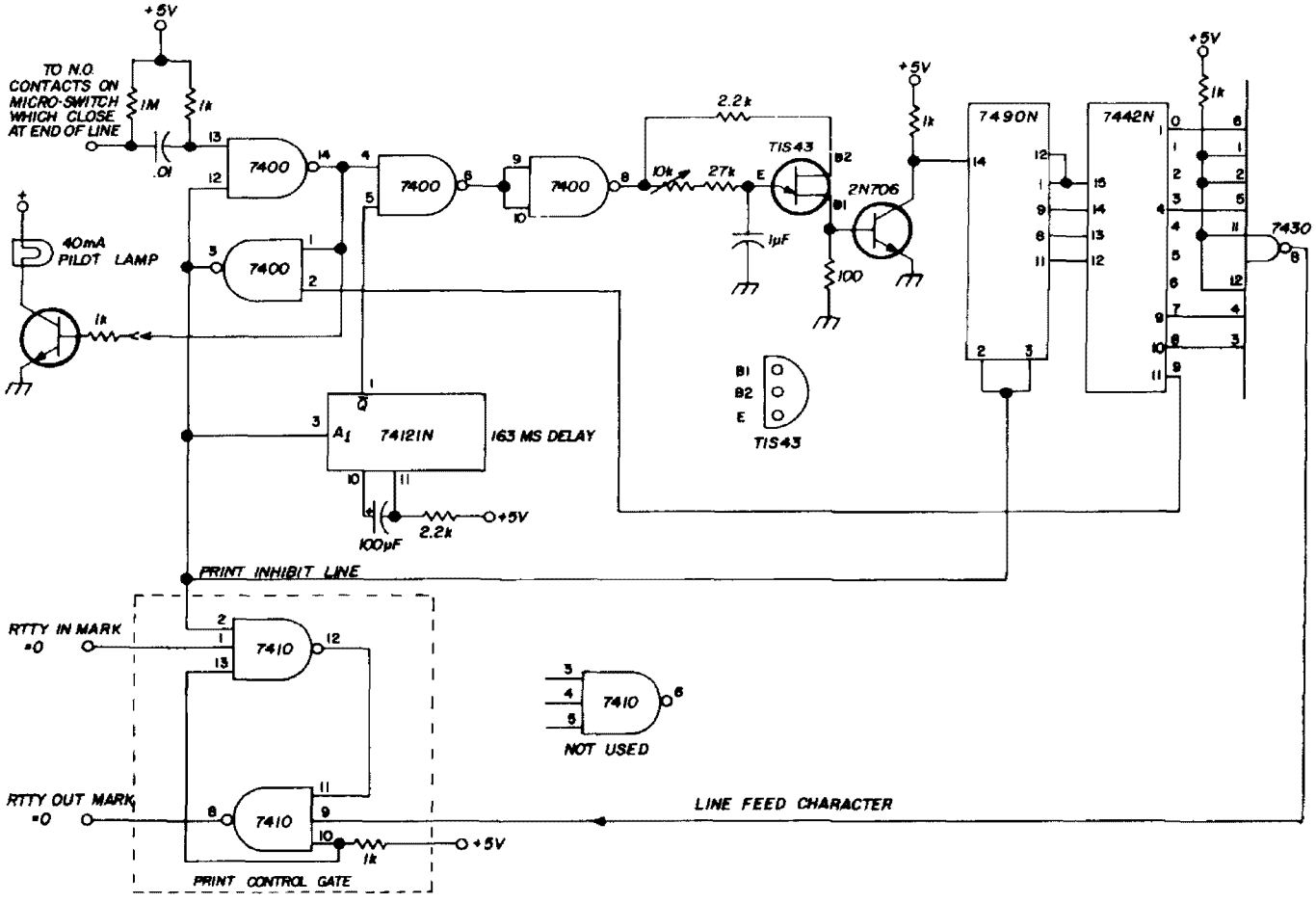


fig. 1. Logic for the RTTY automatic linefeed circuit.

the circuit

This automatic line-feed system uses six TTL integrated circuits and two transistors to generate a line-feed character locally and insert it into the loop. A microswitch is mounted on the frame of the printer so that it is actuated by a projection on the type carriage when the printer reaches the end of a line (see photo). Two gates then act as electronic switches to shutoff print, insert the line-feed character, and restore print. The entire sequence happens very quickly, and only three characters are lost in the copy, even at full machine speed. However, in practice, only one or two characters are lost as few RTTY operators actually type at 60 words per minute (the auto-line feed occurs at full machine speed).

Normally, the output of the TTL NAND gate is at logic 1. The output of the two-input gate will be zero if, and only if, both inputs are logic 1. A 7400 IC is used as a start/stop latch, with the output at pin 14 normally logic zero.

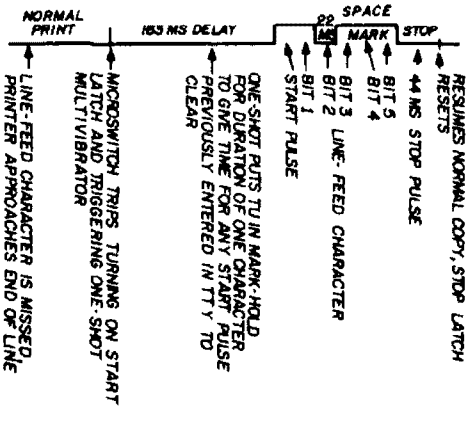


fig. 2. Timing sequence of the automatic line feed.

When the line-feed microswitch closes, the start/stop latch trips, pin 14 goes to logic 1 and pin 3 goes to logic zero, triggering a 74121 monostable multivibrator.

the length of one RTTY *bit*. The 7442 decoder IC is connected to an eight-input 7430 gate so that decimal zero is mark-hold, decimal one is the start pulse, and the remaining bits follow in order

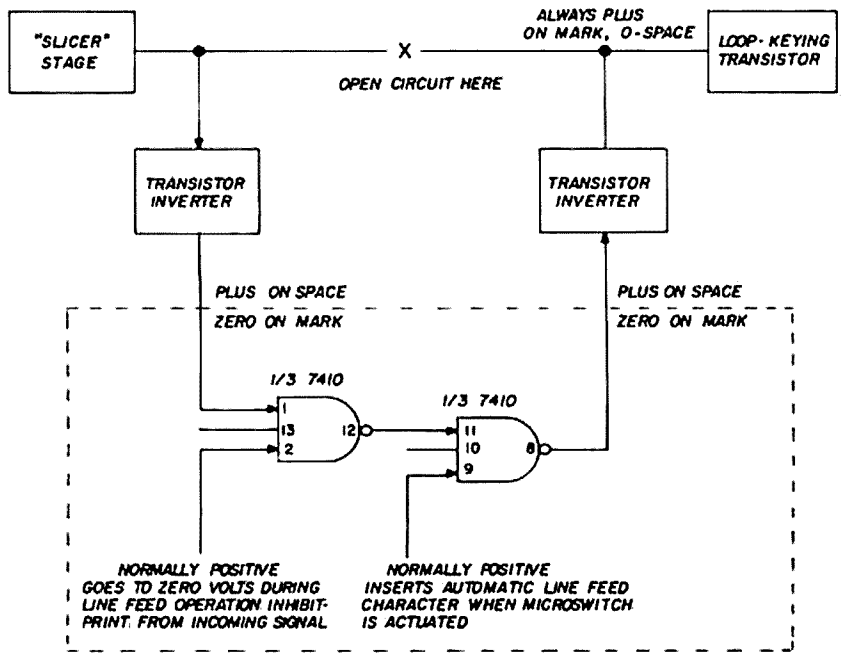


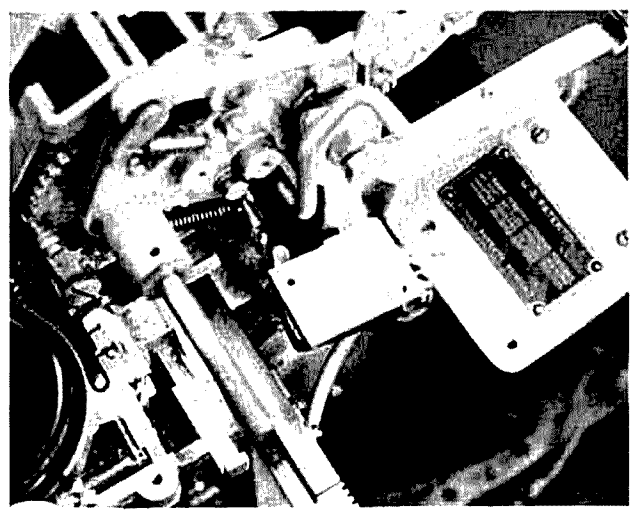
fig. 3. The automatic line feed may be used with most solid-state RTTY demodulators as shown here.

At rest, pin 5 of the 7400 is at logic 1; with pin 14 at logic 1, pin 6 would be at logic zero except that the 74121 has been triggered and the output of the one-shot is low for about 163 milliseconds. This gives the printer time to clear any parts of an incoming signal (print is inhibited by the 7410 gate as soon as the voltage at pin 3 of the 7400 falls to logic zero).

After the one-shot pulse is completed, pin 5 of the 7400 goes high, and since the start latch has tripped, pin 4 is also high, meeting the condition for a logic zero at pin 6. The next 7400 section is used as an inverter and is at logic 1; this is about 4 volts. The 1-μF capacitor in the unijunction emitter circuit starts to charge, and fires the unijunction transistor, generating a sawtooth wave at a frequency of 45.5 Hz. The 2N706 transistor is used as an interface between the unijunction transistor and the 7490 TTL decade counter.

The 7490-7442 counter-decoder combination converts the 45.5-Hz sawtooth to a series of 22-millisecond pulses, each through the sequence until the stop pulse is reached on pins 9 and 10 (decimal seven and eight).

Both pins 9 and 10 of the decimal decoder are connected to the gate, giving a 44-millisecond stop pulse. When the counter reaches decimal nine, pin 11 goes low, immediately resetting the flip-flop to



Line-feed microswitch is actuated by a projection on the type-carriage arm that mounts the margin-bell pawl.

standby and restoring print. The 7490 is simultaneously reset to zero, ready for the next microswitch closure.

The same circuit could be used to generate any RTTY character by proper connections to the 7430 gate. In this circuit only the line-feed character is generated, and it is assumed that the printer is already modified for non-overline; that is, it is only necessary to send a line-feed signal to obtain both carriage return and line feed (kits are available for this*).

installation

The automatic line-feed system described here is compatible with most solid-state transistor or IC RTTY demodulators. However, proper signal polarity must be observed, and the input to the TTL IC must not exceed 5 volts (a zener diode will take care of this). Fig. 3 shows the general interconnection of the automatic line-feed; fig. 4 shows how the ST-3 and ST-4 RTTY demodulators may be adapted, and fig. 5 shows the ST-5.

Power for the automatic line-feed circuit can usually be taken from the demodulator. The logic requires about 80 mA at 5 volts. An LM309K voltage-regulator IC is the most convenient way to obtain the proper voltage from an unregulated demodulator positive supply.

The microswitch is mounted on the teleprinter as shown in the photographs. To prevent ground loops it's necessary to use shielded wiring between the microswitch and the logic. In my unit, I used printed-circuit sockets and phono connectors for easy servicing.

The only adjustment required is that of setting the clock frequency to 45.5 Hz. There are two ways of doing this. One is to temporarily disconnect the reset line at pin 11 of the 7442 decoder; this makes the clock run continuously so its frequency can be measured with a digital frequency counter. If the 10k potentiometer will not reach the proper setting, change the 27k resistor to another value to compensate for wide tolerances in the electrolytic capacitor. If you do not have access to a frequency counter, simply adjust the clock for proper line-feed operation while manually actuating the microswitch.

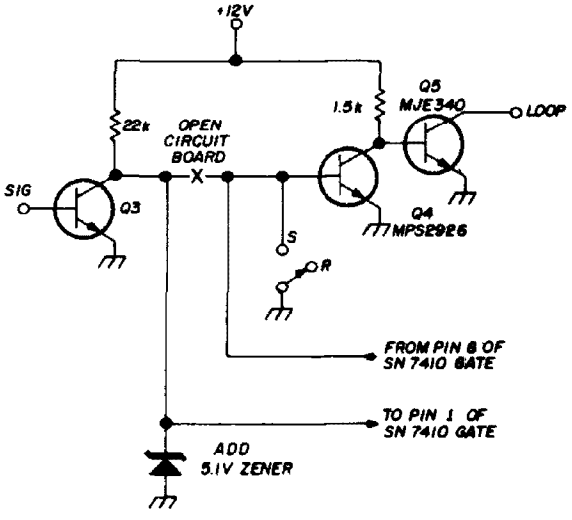


fig. 4. Connecting the automatic line-feed circuit to the ST-3 and ST-4 demodulators.

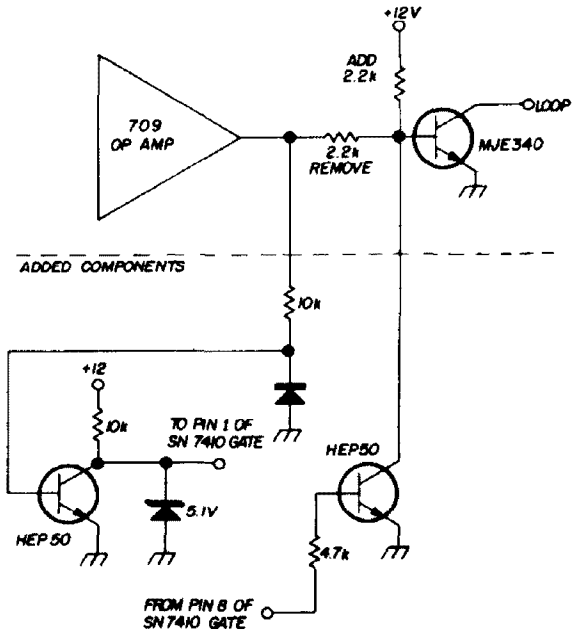


fig. 5. Proper connections for using the automatic line feed with the ST-5 demodulator.

*Modification kits for non-overline are available from several sources, including Typetronics, Box 8873, Fort Lauderdale, Florida.

references

1. Irving M. Hoff, W6FFC, "Simplex Auto CR-LF," *RTTY Journal*, July, 1968.
2. J. Struder, W9RYI, "Modification for the Simplex," *RTTY Journal*, May, 1971.

ham radio

multi-function fm repeater decoder

This multi-function Touchtone-actuated repeater decoder provides over 3500 different functions with a 16-button pad

Bill Bretz, WA6TBC, 110 Field Drive NE, Albuquerque, New Mexico 87112

A repeater usually has some sort of simple control system that requires, perhaps, one tone to turn the repeater on and a timer to shut it off. Some of the more elaborate systems use Touchtone coding which allows the users 12 different functions (or maybe 16) by using the pad from a Touchtone telephone. These functions may be: repeater on; repeater off; tight squelch; open squelch; 450 link on; 450 link off; etc. The system described here is the next step forward. It uses a three-digit combination that provides facilities for up to 810 functions using a ten-button Touchtone pad, up to 1464 functions with a 12-button Touchtone pad, and up to 3616 functions using a 16-button pad.

In developing this system, the intention was not to start off with a full 3616 function decoder, but to build a unit so that functions could be added to the decoder in modules as more and more functions were required. It was decided that it would be easiest to add functions in modular groups of ten. In the decoder described here, up to ten of these modules (100 functions) can be added to the basic decoder. By adding four more integrated circuits, another ten modules can be added (a total of 200 functions). Still another set of ICs will expand the system to 1000 functions.

This article describes a basic 20-function decoder that can easily be expanded to 100 functions. Instructions for further expansion are also included. The decoder consists of many subsystems, some of which may be eliminated or replaced with an existing design. Each subsystem is a distinct entity and may be useful for other purposes apart from this particular decoder.

In addition to the three-digit decoder, which consists of a memory unit and function modules, this system includes a base tone decoder, two different prefiltering systems and a display. Either of the two prefiltering systems may be used to increase the differential tone input range to the tone decoders. The display converts Touchtone coding to binary, and then uses a seven-segment display and binary decoder/driver to display 16 different characters, corresponding to the 16 buttons.

basic touchtone decoder

There are several reasons for using phase-locked loops for tone decoders: their small size, excellent bandwidth, relative immunity to bandwidth with level changes and their ability to decode Touchtone signals when the levels of the two tones differ by as much as 6 dB. However, they are expensive – about $9.00 apiece, and there are eight required for 16 digits, seven for 10 or 12 digits.

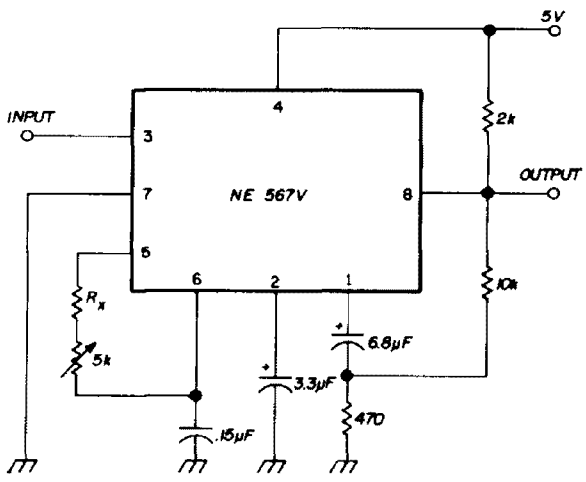
The tone decoder circuit in fig. 1 is typical. Proper resistor values for each of

eight tones are included. Audio, on the order of 200 mV rms is applied to the input of the IC. Each chip has its own RC combination for timing, and will respond only to the tone that it is timed for. Each IC produces a low output when its tone is being received and rests at +5 V. Bandwidth for these decoders with 200 mV rms input and the filtering capacitors shown in the schematic is on the order of 70 Hz.

The decoder, as shown in the schematic, requires the level of the two different tones (differential tone level) to be within 6 dB of each other. This level imposes a fairly stringent requirement on the audio response of the repeater receiver and the repeater users transmitter. Rather than requiring all users to clean up their transmitters, it may be easier to build a pre-filter to increase the differential response of the decoder.

prefilters

Two different prefilter systems were built and tried; both will be described here. One filter system is completely passive. It has two bandpass filters, one for the high group frequencies, the other for low group frequencies. These filters



frequency (Hz)	Rx (ohms)
697	6800
770	6200
852	5100
941	4700
1209	3000
1336	2400
1447	2000
1633	1500

fig. 1. Circuit for the IC tone decoder. One decoder is required for each tone used in the system. The value of Rx determines the Touchtone frequency to which the tone decoder is sensitive.

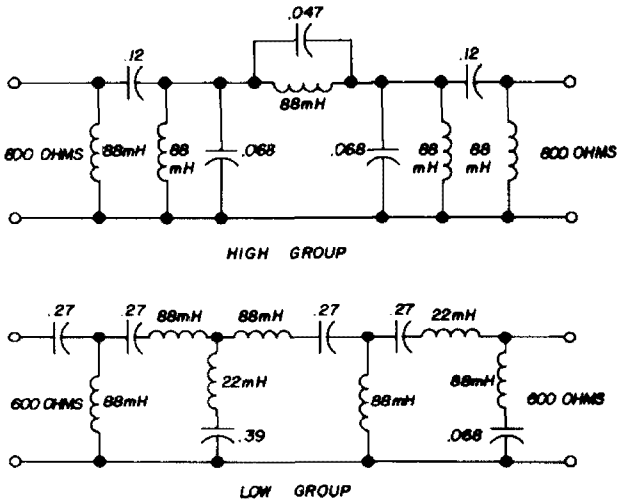


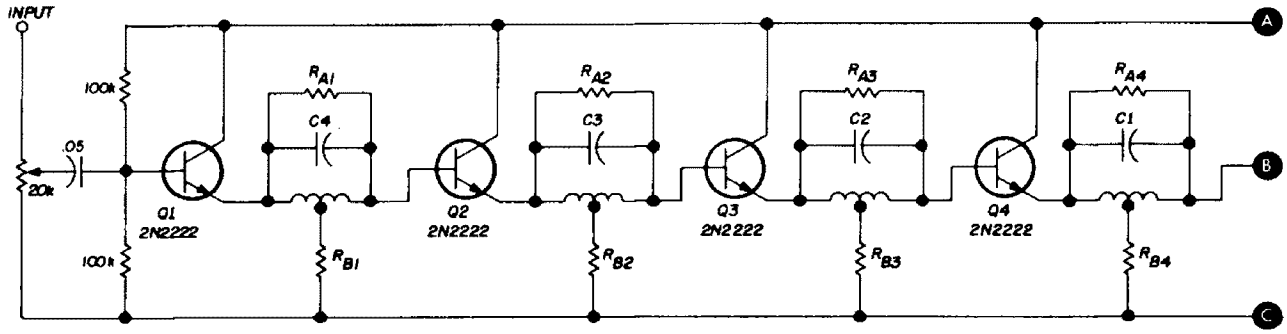
fig. 2. Passive bandpass filters for high- and low-group tones.

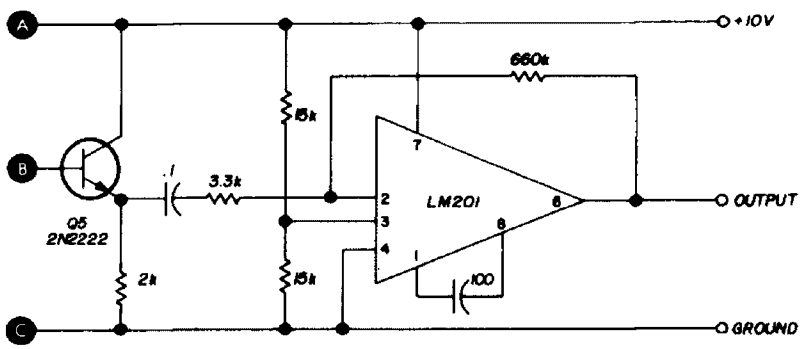
are built from standard 88-mH toroids and off-the-shelf capacitors and require no tuning. Schematics for these two filters are shown in fig. 2.

The other filter system is considerably more elaborate. It basically consists of notch-filters cascaded together with an active limiter at the output and is used for band-reject filtering at the input to the tone decoders. Consequently, all the high-group tones are rejected at the input to the low-group decoder and vice-versa. The schematic for this filter is shown in fig. 3. Tune-up instructions for these filters will be discussed under construction. The passive bandbass filter furnishes an additional 16-dB minimum isolation between the decoders while the active band-reject filters supply a minimum of 20 dB. This allows a total differential tone range of 22 dB for the passive filters and 26 dB for the active filters.

The advantage of the notch-filters over the bandpass filters is that with the notch-filter, the input of the tone decoder is looking at the entire audio spectrum except for the tone that is notched out. This requires that the overall signal-to-noise ratio be better than 6 dB for the decoder to function. With the bandpass filters, the input of the tone decoder looks at a very narrow portion of the spectrum, and the unit will work with very marginal signal-to-noise ratios.

The schematic for the display is shown in fig. 4. Notice that DTL logic is used instead of TTL. This is so that wired-OR can be used at the input to the inverters driving the binary to seven-segment display driver (MSDO47). If TTL logic is to be used, all gates with the exception of U1 must have open-collector outputs and associated pull-up resistors. Another factor to be considered is that the display logic requires positive logic while the Touchtone decoder shown here furnishes negative logic. For operation directly from the decoder, inverters will have to be used between the decoder outputs and the display logic inputs. A Monsanto MSDO47 decoder-driver was chosen; however, an SN7447A could be used in place of the Monsanto unit.





	low group	high group
C1	.68	.22
C2	.5	.18
C3	.39	.15
C4	.33	.1

fig. 3. Active band-reject filter. All inductors are 88-mH toroids. Resistor R_A is between 5600 and 22k ohms, R_B is 1000 to 3000 ohms (see text).

memory unit

The operation of the memory unit requires some explanation. Basically, it consists of a *recognizer* (which recognizes legitimate Touchtone signals and not noise), an *access* system which has access to the memory, a counter to count digits and the memory itself (see fig. 5).

Essentially, the memory unit works as follows. The *access* command is sent to the decoder. This single-digit Touchtone command resets the counter and enables the system to store the next three digits in the memorys. After the next three digits are sent, the unit produces an output to tell the function modules that there is a command in memory (*command enable*), and to shut the access to the memory down until the next access command is sent. It is possible to use single row or file accessing, as is common on the West Coast, with minor circuitry changes.

The four high-group tone decoders are applied to the clock on one set of four lines, while the four low-group tone decoders are applied on another (fig. 6). Each group of lines is *looked* at by a set of exclusive OR gates (U7 and U8). These gates produce an output only when one of the four lines goes low. These two exclusive OR sets are NANDed together and applied to an integrator. This integrator will only start to discharge when *one* of the high-group lines and *one* of the low-group lines go low. This happens only when a legitimate Touchtone command is being sent.

Since this integrator has a 100-ms time constant, the 311 comparator will fire

only if the command is present for more than 100 ms. The output of the comparator is connected to a one-shot so that a pulse is produced when the comparator fires. This 7-ms pulse is produced 100 ms after a button is pushed on the pad. This

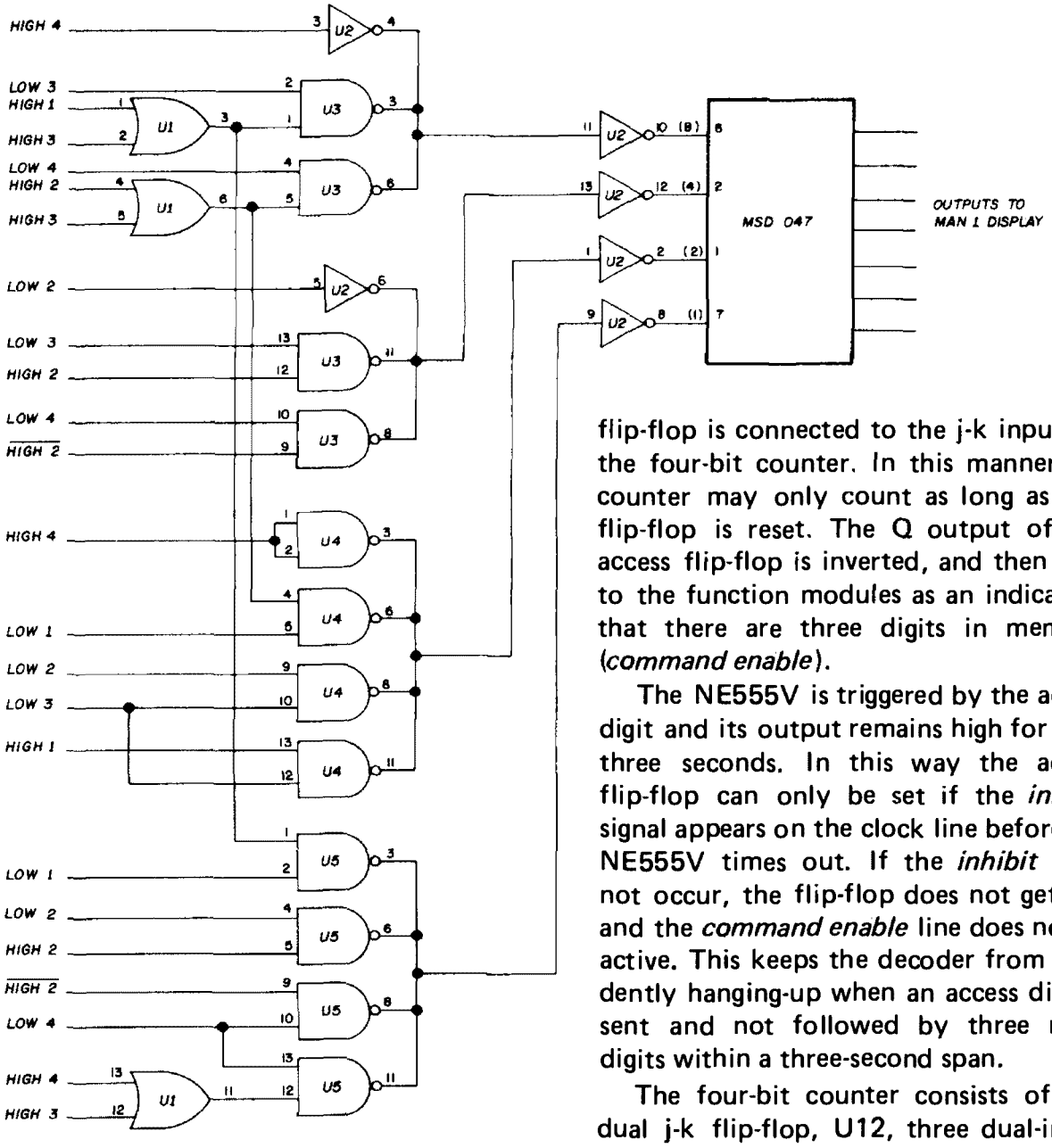


fig. 4. Display logic. U1 is a MC1809P; U2 is a MC837P; and U3, U4 and U5 are MC846Ps.

pulse is fed to one input of a two-input NAND gate while the other input is connected to the input of the integrator. If the command is still present when the one-shot produces its pulse, the output of this gate will produce the *digit* present pulse which drives the four-bit counter.

Gate 2 serves as the access programmer. This NOR gate output goes high when the *access* digit is being sent. The output of the access programmer resets the four-bit counter, resets the access flip-flop (U15), and starts the NE555V timing. The Q output of the access flip-flop is connected to the j-k inputs of the four-bit counter. In this manner the counter may only count as long as this flip-flop is reset. The Q output of the access flip-flop is inverted, and then sent to the function modules as an indication that there are three digits in memory (*command enable*).

The NE555V is triggered by the access digit and its output remains high for only three seconds. In this way the access flip-flop can only be set if the *inhibit* signal appears on the clock line before the NE555V times out. If the *inhibit* does not occur, the flip-flop does not get set, and the *command enable* line does not go active. This keeps the decoder from accidently hanging-up when an access digit is sent and not followed by three more digits within a three-second span.

The four-bit counter consists of one dual j-k flip-flop, U12, three dual-input NOR gates and a two-input NAND gate (fig. 7.). The counter is programmed so that when it is reset, output 1 is high, outputs 2 and 3 are low, and output 4 is high. Output 1 occurs when the counter is reset by the access command. Output 1 is connected to the first digit memory (1st digit memory enable). This high input to these two memory units allows them to look at their input and remember what it was when this line goes low again.

If a *digit present* pulse appears at the input to the counter, the counter will advance, making output 1 low and output 2 high with no change in outputs 3 or 4.

Thus, the digit that was present when the digit *present* pulse occurred is stored in the first digit memory, and the second digit memory is ready to accept a digit. When the next digit is sent, another *digit present* pulse is produced, locking this digit in the second digit memory and allowing the third digit memory to accept

For single rank or file accessing (i. e., single-tone produced by pushing two adjacent buttons on the pad), the following addition must be made. For single-tone, low-group accessing, a four-input NAND gate is installed between gate 2 and the high-group tone decoder inputs. The output of this added gate is connected to pin 12 of gate 2, and the inputs are connected to the high-group tone decoder inputs. This gate functions as a NOR gate in this configuration, and gate 2 will now

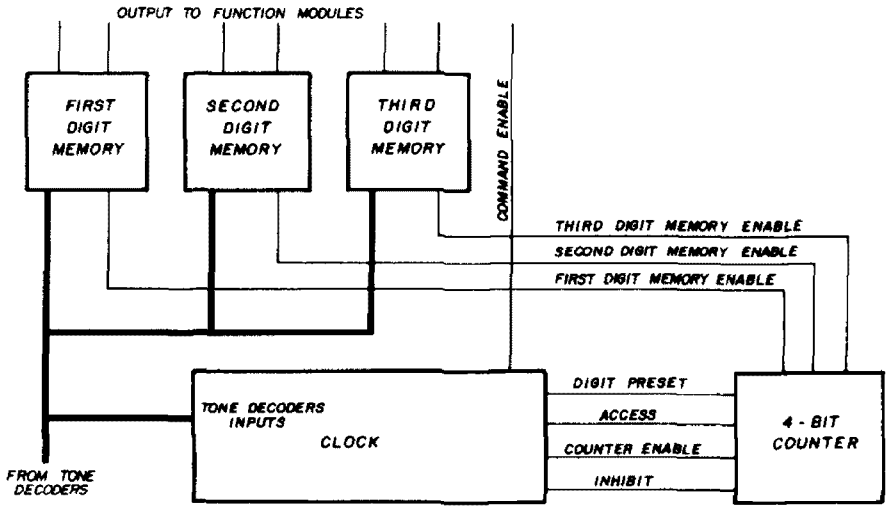


fig. 5. Block diagram of the mamory.

a digit. The same thing happens when the third digit is sent, except that output 4 now goes low which sets the *access* flip-flop so that no more *digit* present pulses may get to the four-bit counter unless it is accessed again. The access flip-flop can be set only if the NE555V has not timed out. Therefore, output 4 of the counter must occur within 3 seconds after *access* or the access flip-flop will stay reset.

The memory consists of two SN7475 quad latches for each digit (**fig. 8**). The output of the memory to the function modules is from the Q outputs. The unit is so constructed that resending the *access command* at any time resets the four-bit counter and reaccesses the counter input, as well as restarting the NE555V. Thus, the *access command* can be sent at any time to restart the decoder. By the same token, this *access digit* cannot be used as a digit in the three-digit command.

produce an output only when none of the high-group tones are present and when the low-group tone, to which the other input of the access programmer is strapped, is present. For single-tone, high-group accessing, put the added gate between the low-group decoder inputs and pin 11 of gate 2.

function modules

The function modules are built in groups of ten. One pair is shown in **fig. 9.** Each pair of functions are electrically the same, except for programming. Each group of ten functions is prefaced by hex-inverters on the memory output lines for fan-out; this is why the number 10 was chosen for a group of functions. Each function module is an eight-input NAND gate, followed by a two-input NAND gate. One of the eight inputs is permanently tied high, so the gate functions as a seven-input gate.

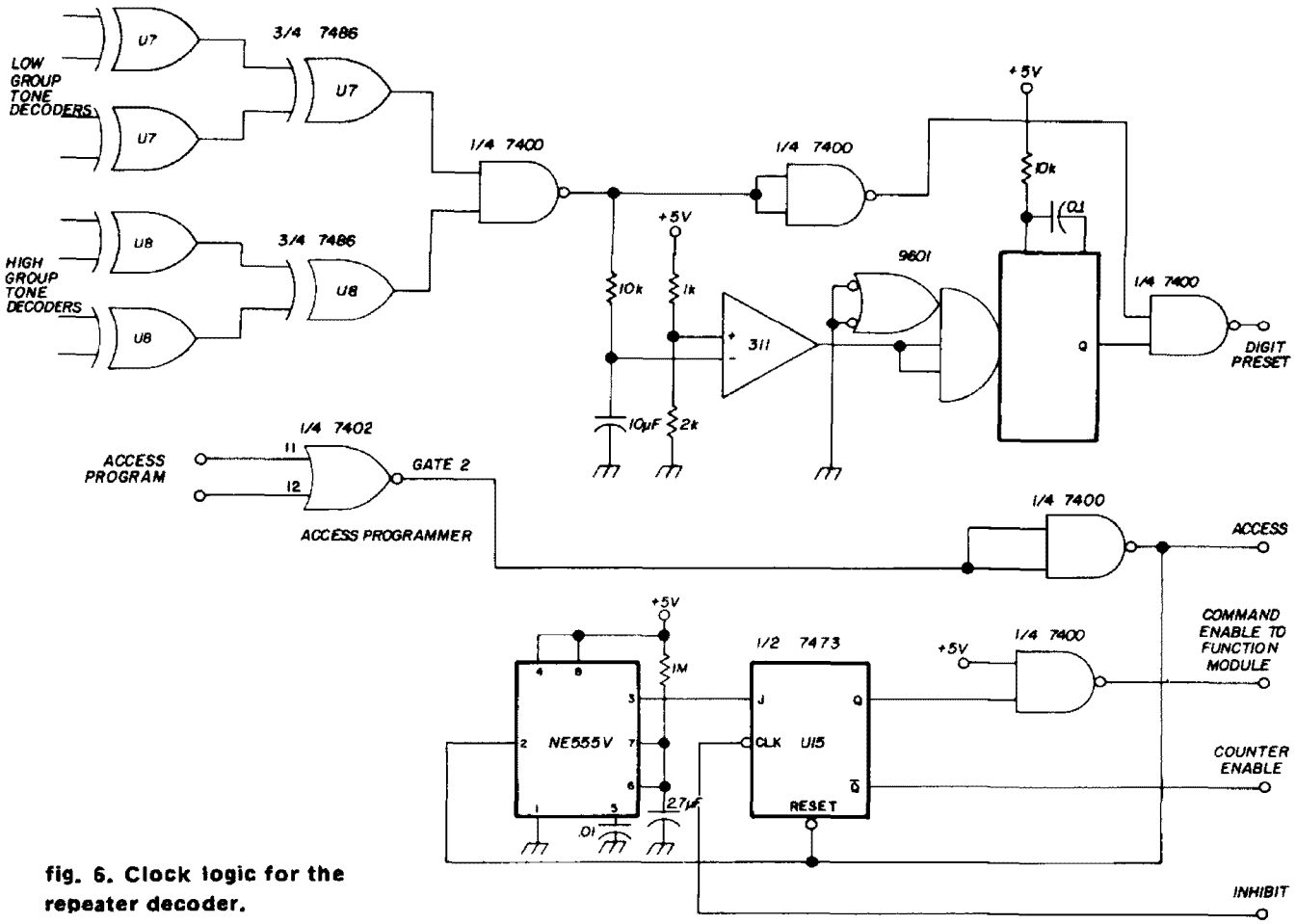


fig. 6. Clock logic for the repeater decoder.

Programming is accomplished by tieing two of the inputs to the right coding for each digit. The seventh input is tied to the *command enable* line. In this way, if the right three digits are in memory, and the *command enable* is active, the eight-input gate will produce a low output. The output of this gate is applied to one input of a two-input NAND gate. By strapping with the next adjacent command, it is possible to use one three-digit combination to turn something on, and another three-digit combination to turn something off. If latching output such as this is desired, then strap 2 and 6 together, and 3 and 5 together. Then command 1 will remain high until command 2 is sent, and command 2 will remain high until command 1 is sent. If momentary outputs are desired, i. e., the output only stays high until the next time the *access digit* is sent, strap 1 and 2 together, and 4 and 5 together.

construction

Printed-circuit boards were used for the tone decoders, each of the pre-filters, the memory unit and the function modules. Three of the display logic assemblies were built on one wire wrap card, due to the high density. The tone decoder board has provisions for 8 circuits. Do not try

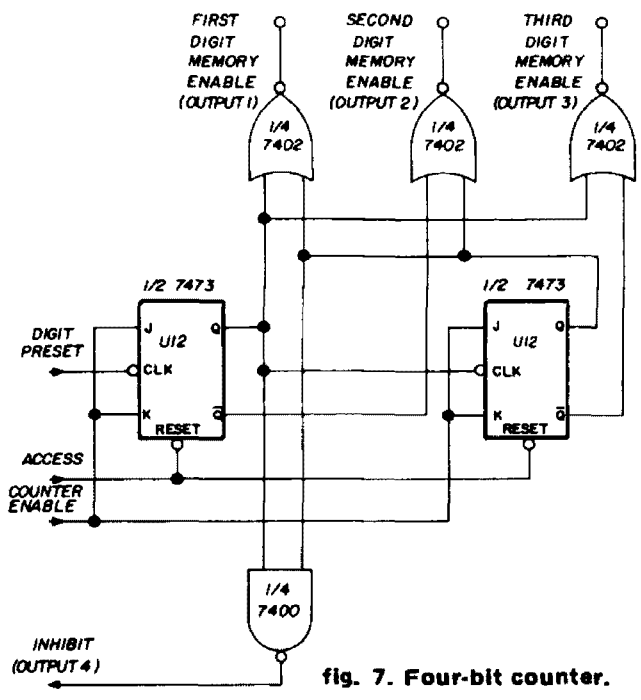


fig. 7. Four-bit counter.

to substitute values for any of the capacitors, otherwise you may have bandwidth and/or temperature stability problems.

The best way to set the center frequency of each decoder is to connect a counter to pin 5 of the decoder being aligned and adjust the pot for the proper center frequency.

For use without prefiltering, all inputs to the decoders are tied together and a 1 μF coupling capacitor is used to connect this common input to the audio input. When using either of the prefilters, the audio inputs of the four low-group decoders are tied together and called the low-group audio input; the four

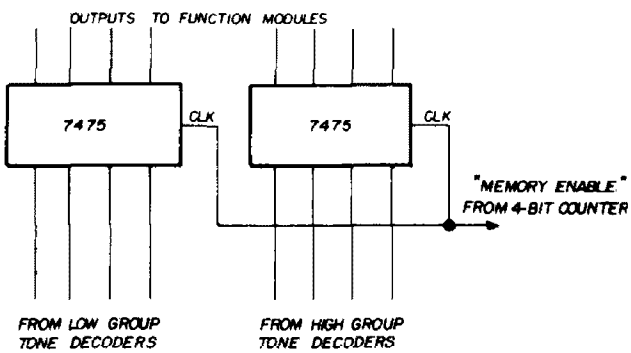


fig. 8. Memory (typical of three).

high-group decoder audio inputs are tied together and called the high-group audio input. When using the passive prefilters, both the low-group audio input and the high-group audio input will require a 680-ohm resistor to ground to normalize its impedance to 600 ohms, and each of these inputs is connected directly to the output of its filter. When using the active prefilter, both the low-group audio input and the high-group audio input are connected to the output of its filter through a 1-μF capacitor.

When building the passive prefilter shown in fig. 2 it is essential that the capacitor values be exactly as shown. There is no tuneup necessary for these units. However, the input and output impedances are critical. Each filter must see 600 ohms at both input and output. Connect the filters as shown in fig. 10. If a 600-ohm line is available for use, the 600-ohm resistor can be removed. If all that is available is a speaker line, use an output transformer backwards to jack the impedance up.

The active prefilters are built on two identical PC cards, one for the high group, the other for the low group. It is easier to begin tuning this filter before it is constructed. Temporarily connect C1 across an 88-mH toroid. Install the combination in the test fixture shown in fig. 11. Measure the resonance point of the combination. This is indicated by a peak on the voltmeter. Chances are, the peak will not correspond with the needed frequency. For low-group, C1 should produce resonance at 697 Hz, C2 at 770 Hz, C3 at 852 Hz, and C4 at 941 Hz. For high-group, C1=1209 Hz, C2=1336 Hz, C3=1447 Hz and C4=1633 Hz.

In all cases, the resonance should be lower in frequency than desired with the capacitors called for in fig. 3. To tune the combination to the proper frequency, start removing turns from the toroid. Remove turns equally from both windings. After tuning, install this LC combination on the circuit board. Connect C2 across another toroid, install in the jig, and continue as above, tuning the combination for 770 Hz. Do the same for C3 and C4.

Temporarily install 2000-ohm resistors at RB2, RB3 and RB4. Do not install RA1, RA2, RA3 or RA4 at this time. Connect an audio oscillator and counter to the input of the filter board. Check the resonant frequency and notch depth of first tuned circuit by connecting an ac voltmeter to the emitter of Q2. Use a decade resistance box for RB1 to determine the correct resistance for the deepest notch (typically 50 dB). It may be necessary to add or subtract turns from the toroid to finalize the frequency of the notch.

Install a fixed resistor for RB1 and connect the decade box where RA1 should be. Determine the value of RA1 required to broaden the notch to ±10 to 15 Hz. This should bring the notch depth to about 20 dB. Permanently install RA1, and continue on to the next frequency by removing the temporary RB2 and replac-

ing it with the decade box and moving the voltmeter to the emitter of Q3.

access programming

Pick one of the digits on the Touch-tone pad for access. Let's use Ø as an example. Referring to table 1, digit Ø is low-group 4 and high group 2. Run jumpers from high-group tone decoder 2 (1336) and low-group tone decoder 4 (941) to the two-input NOR gate used for access. This is all that is required for access programming.

To program the decoder for specific digit combinations, pick three digits (except for the *access* digit). Let's use 159. Referring to table 1, digit 1 is high-group 1, low-group 1. Run jumpers from the first digit memory outputs *high 11* (1H1) and *low 1* (1L1) to two of the inputs on the eight-input gate. This programs first digit one.

Referring again to table 1, digit 5 is high-group 2 and low-group 2. Run jumpers from second digit memory outputs *high 2* (2H2) and *low 2* (2L2) to two more of the inputs of the eight-input gate. This programs second digit 5. Digit

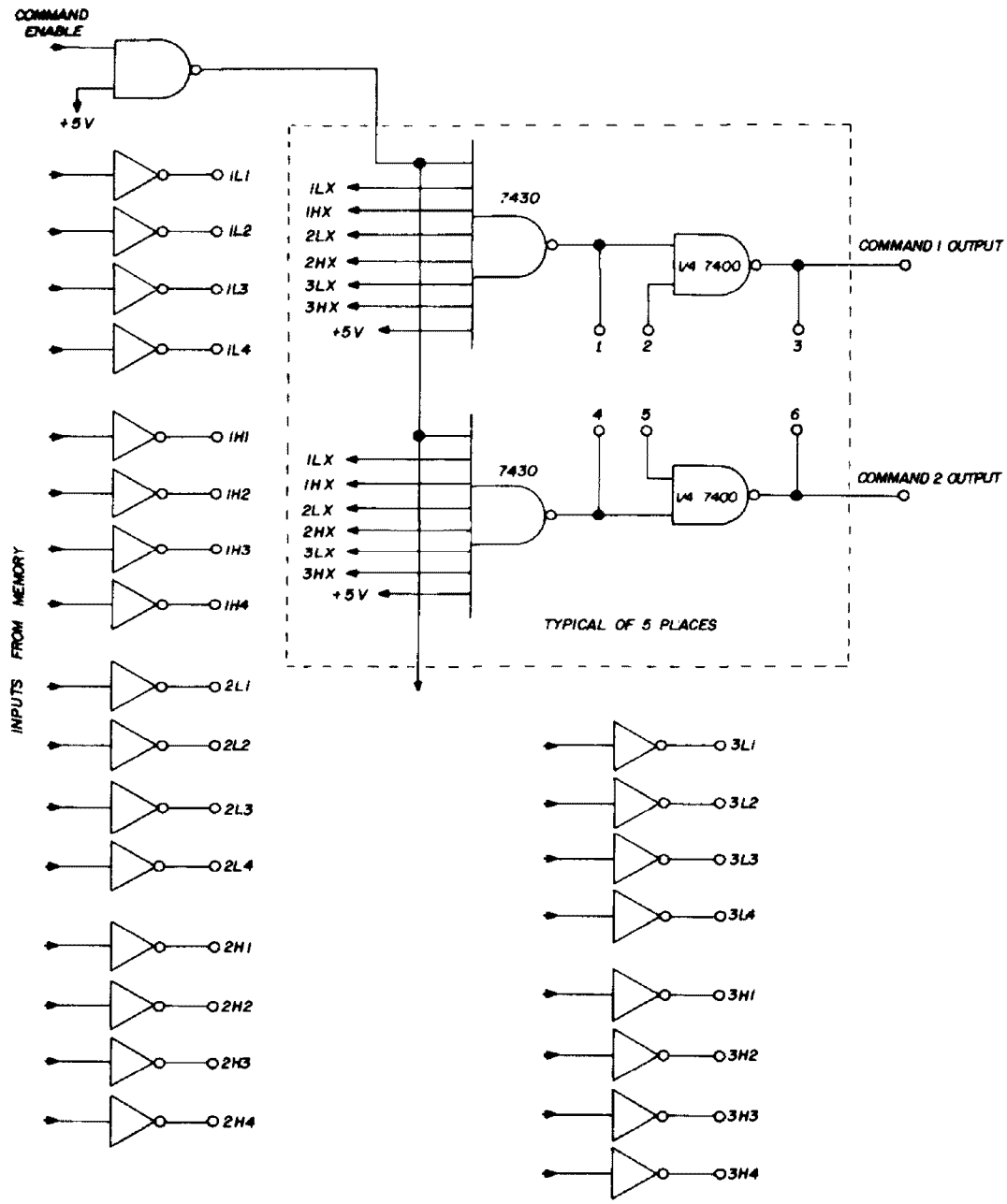


fig. 9. Function module board.

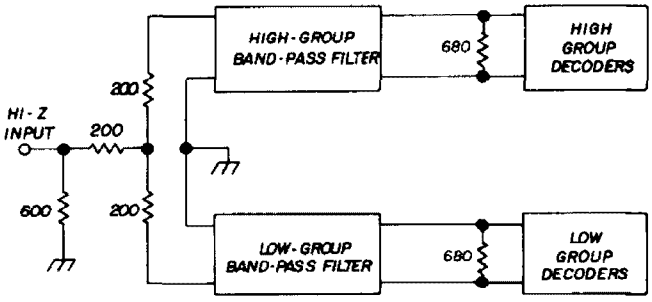


fig. 10. Wiring of the bandpass filters.

9 is high-group 3, and low-group 3. Run jumpers from third digit memory outputs *high 3* (3H3) and *low 3* (3L3) to the last two inputs of the eight-input gate. This programs third digit 9. Now, whenever the access digit is sent followed by 159, the output of this module will go high. Note that if *access digit* and then 519, 951, 195 or 915 or any other combination of these 3 digits except 159 is sent, nothing will happen.

complete system

The complete unit consists of the sub-assemblies installed together in a box. It was decided to make the unit rack-mountable. In keeping with the easy-to-expand philosophy, it was decided to mount everything except the function modules in one chassis. The function modules were mounted in a separate chassis so that more and more chassis could be added as expansion is required. On the initial model the tone decoders, prefilters, memory, three-digit display and power supply are mounted behind one 1¾-inch rack panel.

The three-digit display consists of three of the circuits shown in fig. 4 and are connected to the NOT outputs of the quad latches on the memory board. Therefore, the display displays the three digits in memory, even if the three digits are not a command. This chassis has a BNC connector for audio input, a fuse, a 9-pin connector for power output and a 25-pin connector for logic output.

Two of the ten-function module boards are mounted behind another 1¾-inch panel. This gives 20 functions per 1¾-inch rack space. This chassis has two 25-pin connectors, for logic in from the memory unit and logic out to the next function-module unit and a 9-pin connector for power.

Expanding the decoder is simple. Up to ten groups of ten function modules can be added in parallel with each other with no other consideration. Adding another set of ten groups can be accomplished by adding inverters between the NOT outputs of the memory and the NOT output of the *command enable* flip-flop and driving this set of ten groups from these inverters. More sets of ten groups can be added as above, until there are a total of ten sets being driven by the NOT outputs.

table 1. Digit information for programming the decoder.

low group	high group			
	1	2	3	4
1	1	2	3	Fo
2	4	5	6	F
3	7	8	9	I
4	*	ø	#	P

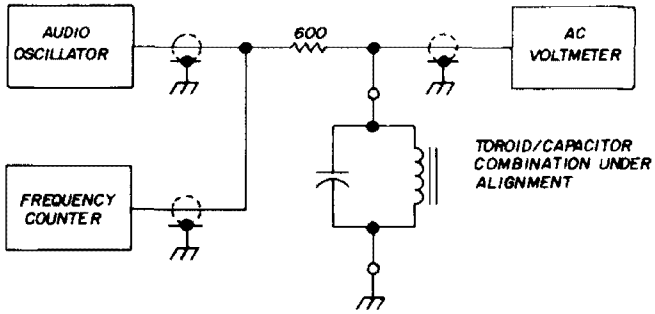


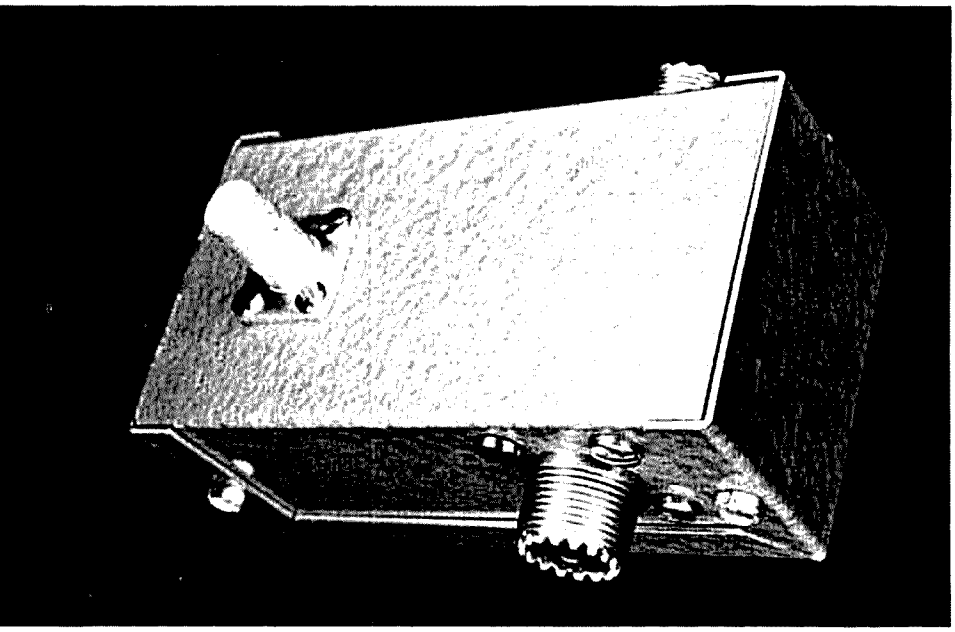
fig. 11. Test jig for tuning up the active band-reject filters.

summary

As of this writing, three of these decoders have been built. All are in use in different Northern California repeater systems. The unit appears to be very reliable and allows for great versatility in the system.

My profound thanks to Dave Bradley, K6AMA, who did much of the preliminary breadboarding, and to Lance Ginner, K6GSJ, who offered so much skepticism about the unit working, that he forced me to actually build it.

ham *radio*



rf sampling network — the Milli-tap

This unit allows accurate measurements of rf voltage, vswr, frequency, spectrum and power without materially affecting the vswr on the line

Howard F. Sheperd, Jr., W6QJW, Del Mar, California 92014

Most amateurs use coax transmission line to deliver their transmitter output to the antenna. While there are many test units that will give an indication of what goes on inside the line — such as directional wattmeters, impedance and vswr bridges — high-quality instruments are priced beyond the operator's pocketbook. Those of lesser quality are not only inaccurate but can themselves upset the transmission line circuit to compound errors.[1;2] In an effort to solve this dilemma, the Milli-tap was developed and its specifications disclose the advantages to be gained.

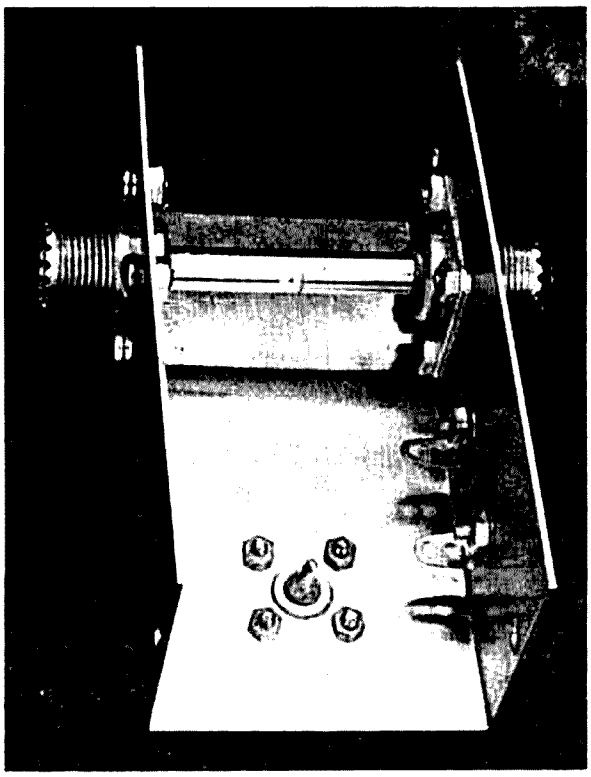
First, its vswr is less than 1.08:1. Hence, it can be put into a line without materially degrading the system. The attenuation to signals in the coax line is virtually zero (the major loss is in the chassis connectors themselves), and the attachment of readout instruments such as vtvms, oscilloscopes or whatever have no effect on the line operation. The test signal measurement point provides very close to 1/1000th of the coax line signal level across 50 ohms and hence duplicates in miniature the voltage in the line itself. With the recommended parts it can be used with transmitters running the full legal maximum power on CW, a-m or ssb.

construction

Construction is made easy by using an LMB "Tite-Fit" aluminum chassis, 4 x 2 1/8 x 1 5/8 inches in size. All of the circuit is constructed in one half of the chassis with the other forming the cover. The drilling layout appears in fig. 1. The heart of the unit is the air coax section, and this is where to begin. Each of the female uhf coax connectors (SO-239) is

altered by carefully sawing off the terminal and then filing it to a length of about 1/32 inch. The photograph shows how each will appear when finished. Apply a touch of solder to the stud end, wipe clean while hot and mount in the chassis holes with 6-32 machine screws and nuts. Tighten lightly as the screws will be removed later.

At this point, install the BNC female fitting which will provide the test signal point. In order to stay as close as possible to an impedance of 50 ohms, the brass center conductor rod should be machined from 5/16-inch stock to a diameter of 0.270 inch and a length of 1-16/32 inches. If you do not wish to be a purist, quarter-inch rod can be used, although the overall accuracy will suffer. The center of each rod end is drilled to a depth of 1/16 inch using a number 30 drill. A hole is also drilled in the center of its length with a number 60 drill to a depth of 3/16 inch. Use the edge of a thin file to create a flat notch centered on this small hole to a width of 1/8 inch and a depth of 1/16 inch.



Partially completed assembly with the air-coax center conductor in place.

Heat one end of the rod while holding it vertical and melt enough solder into the end hole to fill it almost to the top. After cooling, do the same to the other end. File away any excess, clean out each hole with the number 30 drill to a depth of 1/32 inch.

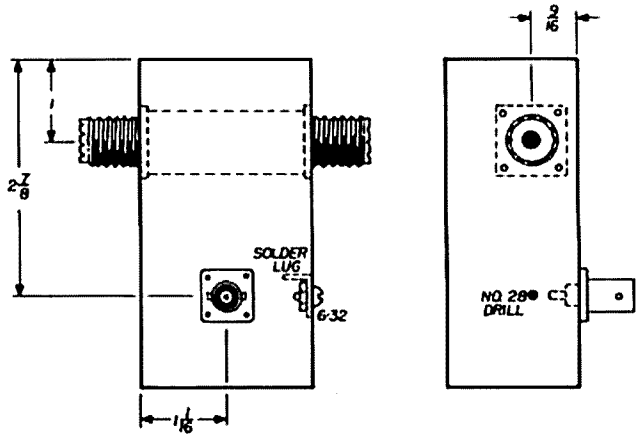


fig. 1. Drilling template for the Millitap.

Next prepare R1 by cutting one lead to 1/8 inch and the other to 3/16 inch. Tin both ends. Heat the center conductor rod with a heavy soldering tool on the opposite side of the center hole and place a very small amount of solder in the small hole. Keep the rod warm and insert the 1/8-inch end of the R1 into the hole until the body of the resistor fits all the way down in the slot. Cool it well without moving the resitor or rod. Clean off any excess solder from the area. Place the rod between the coax plug terminal stubs by first clipping one rod end over the terminal stub and then the other. You should be able to preliminarily position it so R1 is parallel to the back of the chassis and projects toward the test signal terminal. Again heat each end of the center connector rod until it slips over a coax terminal. Some amount of solder will be forced out of the joint as the chassis box edges are clamped toward each other. Clean it away using a knife or file. The photograph shows the assembly complete to this point.

outer conductor

The outer conductor is next cut from

0.003-inch brass shim stock using lightweight tinsnips to form a rectangle 1-26/32 inches wide by 1 31/32 inches long. Carefully drill a hole 1/4 inch from a short edge, centered on the sheet, with

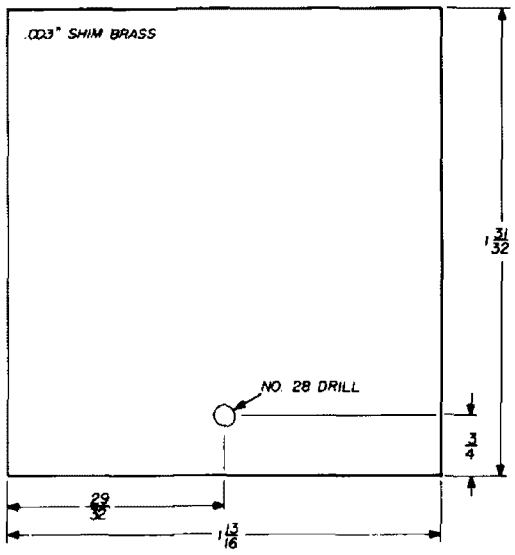


fig. 2. Drilling and cutting template for the outer conductor of the air coax section.

a number 26 drill. See fig. 2 for detail. Using a piece of half-inch dowel or a half-inch drill shank, form the conductor sheet into a cylindrical shape. Remove the nuts and bolts holding the female uhf coax connectors and, after slipping R1 through the hole in the conductor sheet, form it around the shoulders of the connectors. Cut four pieces of solid number 24 to 28 AWG tinned bare wire and pass them around the line assembly, twisting the ends with long nose pliers until the conductor sheet edges meet smoothly. The chassis should now appear as in the third photograph (the upper soldering lug was not used in the final version).

After rotating the uhf connectors so that their holes and the chassis holes align, the outer conductor is finished by applying solder to the point where each of the four retaining wires crosses the gap. Put a generous drop of solder on the top of a hot iron and apply it to the bottom of the outer conductor so that solder will not flow inside the line. After inspecting to see that each retaining wire is secured to each edge of the outer conductor sheet, use small diagonal pliers to cut the wires on each side of the joints. Replace the nuts and bolts securing the uhf connectors, but this time bend two solder lugs into right angles so that when placed under the top nuts on each connector the flat side of the lugs will be right at the surface of the outer conductor. Solder each lug after the bolts and nuts have been tightened. The air coax section is now finished. Next install the attenuator network R2, R3 and R4 as shown in fig. 3 and the photograph.

The unit is now ready for final testing. This is accomplished by using a voltage source of from 100 to 300 Vdc or a 60 Hz ac voltage which is connected between the inner conductor of one of the uhf plugs and the chassis. Be certain to exercise every precaution to avoid personal contact with this high voltage. Measure this source voltage carefully using the most accurate voltmeter you have available. Next measure the voltage across the BNC test signal point. It should be 1/1000th of the applied voltage (0.1 volt for 100 volts, 0.3 volt for 300 volts, etc.). As resistor tolerances are rather wide and the effects of heating during soldering are unpredictable, an exact

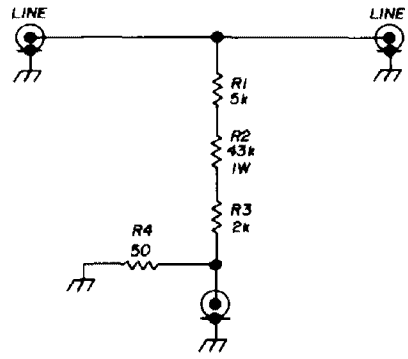


fig. 3. The voltage divider network. All resistors are ½ watt except R2 which is rated at 1 watt.

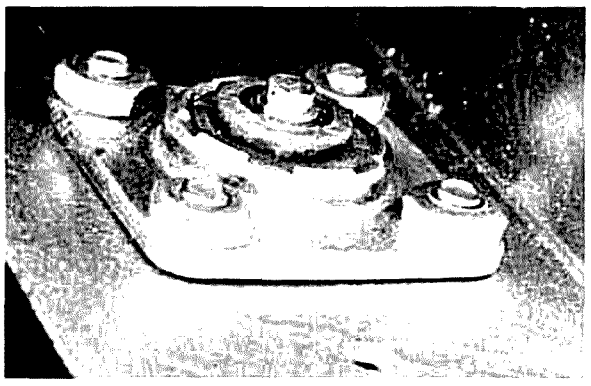
1000 to 1 division may not occur; however, tests with several of these units have shown very close results.

If you are satisfied as to the accuracy of your voltmeter but an exact 1000 to 1

ratio does not exist, change the value of R3 up or down as needed to get the exact division. It should be noted that while this test is carried out with dc or 60-Hz ac, it can be made at any rf frequency and the values of the attenuator changed to obtain even greater accuracy at your frequency of interest. Tests have shown however, that for ranges up to 56 MHz, the dc or ac check is generally sufficient. After testing fit and bolt on the cover portion of the chassis.

operation

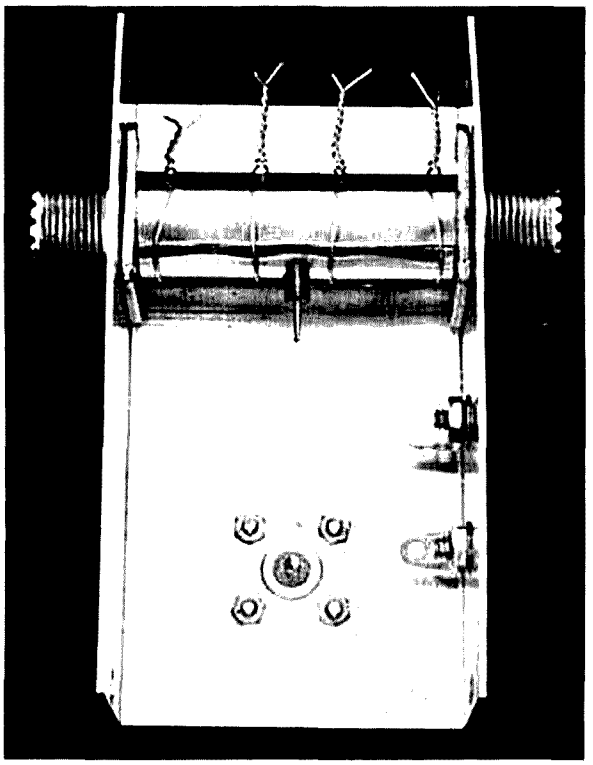
To use the Milli-tap, simply insert it in your coax line at any convenient point near your transmitter. If you wish to check power, load into a 50-ohm non-reactive dummy load and read the voltage at the test signal point on a vtvm with an rf Probe or on an oscilloscope. Most vtvms are calibrated with rms values. Remember your scope will show peak values. Apply the formula $P=(1000E)^2/50$, where E is the voltage measured at the output tap, to determine the power being transmitted in the line. Please note that this power test is accurate only if your antenna feed point is truly 50 ohm non-reactive in nature *and* you are using true 50-ohm coax or if you are feeding a 50-ohm dummy load through 50-ohm coax. If the foregoing is



The modified SO-239.

not true, then the actual power cannot be determined without first determining the impedance existing at the point where the Milli-tap is inserted into the transmission line. However, it can be used for relative power measurements and for other purposes as noted below.

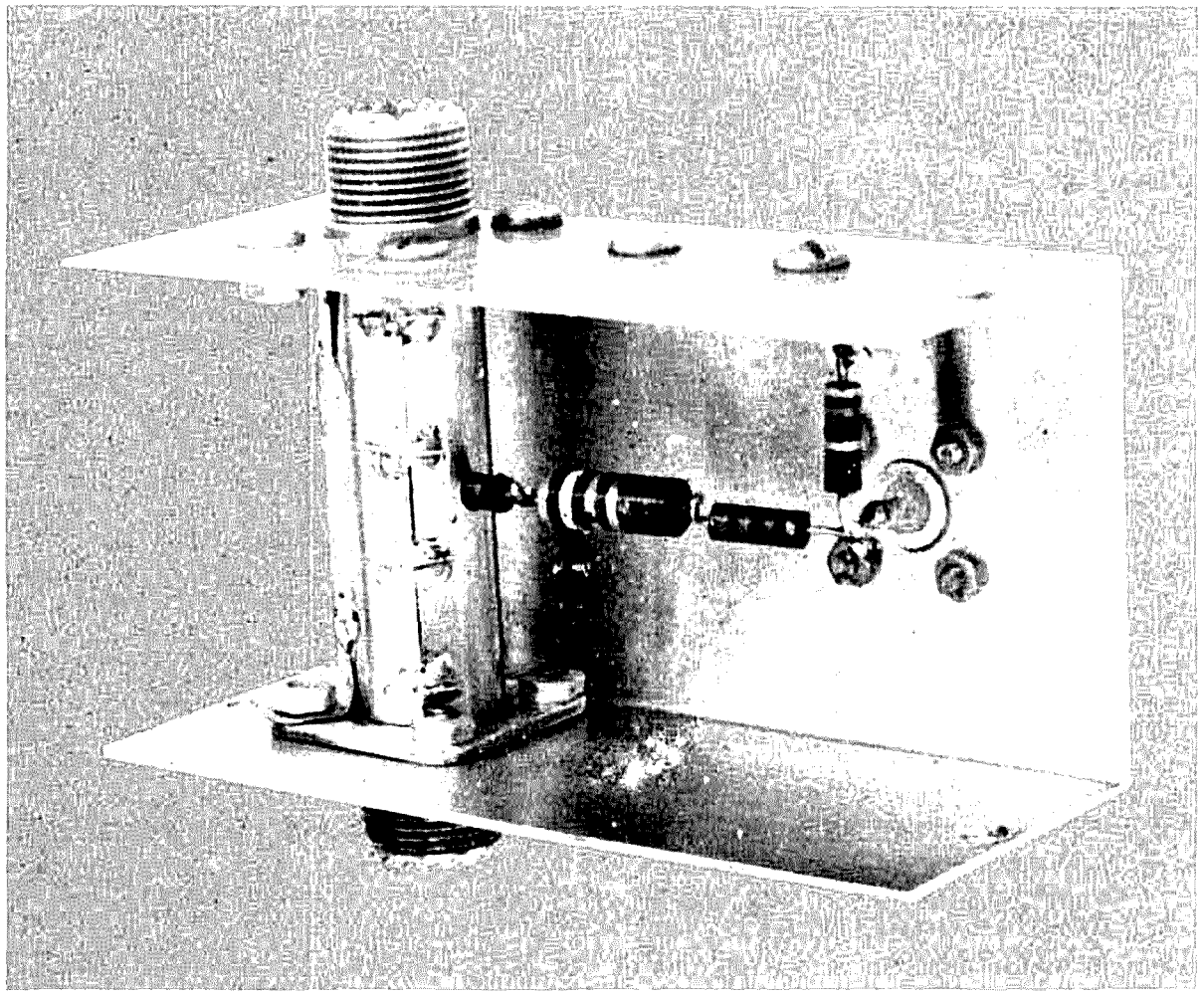
The Milli-tap also furnishes a means of determining vswr by use of two one-eighth wavelength sections of coax line



Construction stage in which the air-coax outer conductor is being formed and held in place by twisted wires. Note resistor R1 which is led through the hole drilled in the sheet brass.

for the frequency to be used which are inserted in the antenna transmission line as shown in fig. 4A. The Milli-tap is then inserted in turn into the three junctions that result, and voltage readings are recorded with a constant power applied to the line in the manner described by Fisk.[3]

Another less sophisticated approach is to utilize a single quarter-wave line section at the frequency of interest, per fig. 4B. Apply some convenient level of transmitter carrier power and read the voltage at signal test point A. Next remove power and move the Milli-tap to the opposite end of the quarter-wavelength section. Restore the same amount of power and again read the voltage. Obviously, if the voltage readings are essentially the same the system being measured is close to a true 1-to-1 vswr. If not, adjustments can be made to the antenna with further



The completed Milli-tap.

measurements to minimize the voltage difference.

An hf oscilloscope can also be connected to the test signal point in order to observe the wave shapes produced by the transmitter or a spectrum analyzer can be tapped in to examine carrier suppression, sidebands, harmonic or spurious signals. Likewise, a sensitive frequency counter can be so connected to monitor carrier frequency. Less sensitive counters may be used by utilizing an amplifier to raise the level of the test signal to that required by the instrument. Many other test applications are possible and provided the maximum voltage in the coax line itself does not exceed 325 volts rms (which represents over 2000 watts PEP in a nonreactive 50-ohm line). Other test procedures will suggest themselves to the reader, and due to its simplicty, the Milli-tap should give a lifetime of performance.

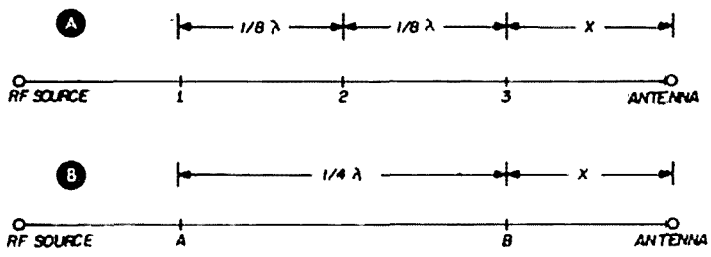


fig. 4. Vswr measurement setup. Junctions 1, 2, 3, A and B are fitted with plug connectors to match the Milli-tap and are bridged with barrel coax connectors.

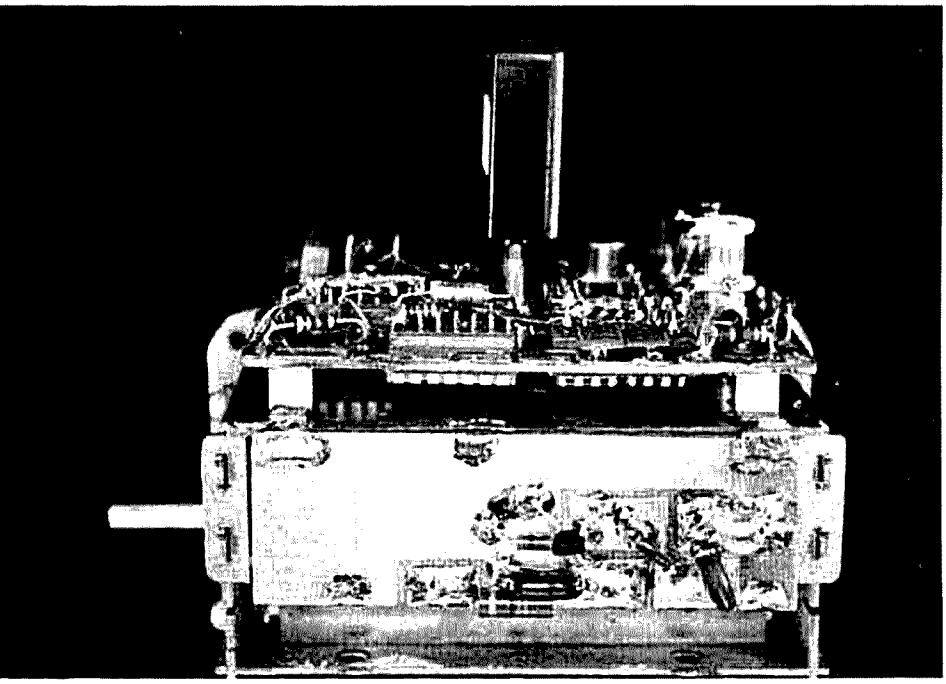
references

1. Louis D. Breetz, W3KDZ, "Possible Errors in VSWR Measurement," *QST*, November, 1959, page 22.

2. Jerry Hall, K1PLP, "Accuracy of SWR Measurements," *QST*, November, 1964, page 50.

3. Jim Fisk, WA6BSO, "Simplified RF Impedance and Power Measurements," *73*, January, 1966, page 20.

ham radio



tunable six- and ten-meter phase-locked loop

This novel circuit is based on the use of a D-type flip-flop

Kenneth W. Robbins, W1KNI, Sudbury, Massachusetts 01776

Dual or triple conversion in vhf receivers is a common method of obtaining the adequate frequency stability required in modern communications systems. The crystal-controlled first oscillator heterodynes incoming signals to a lower band where tunable oscillator stability is acceptable. Single conversion using a high i-f such as 9 MHz and a tunable heterodyned first oscillator is another approach. A frequency-synthesized first oscillator is yet a third technique, but currently too costly for ham use. Described here is an inexpensive version of a circuit developed by Motorola.[1]

This circuit uses the unusual properties of a type-D flip-flop operating as a mixer, phase locking a vhf oscillator to a stable broadcast-band vfo. In this unit a vfo, tunable from 0.5 to 1.5 MHz, has a locked oscillator tracking it from 48.4 to 49.4 MHz. This latter signal mixes with 50- to 51-MHz signals and heterodynes them into a fixed i-f at 1.6 MHz (a solid state car radio).

Vhf stability is the *same* as the vfo which can be made very stable and easily checked against broadcast stations. There is nothing special about the various frequencies; these can even be chosen to obtain conversion to 20 meters.

D-type flip-flop

Since the most intriguing item in this circuit is a D-type flip-flop, some parameter tests were run on a common SN7474N. It performed very well for clock frequencies up to 9 MHz, and with the D input signal, ran up to about 80 MHz. Normal operation causes an odd sort of square wave to be developed at the Q and $\overline{Q}$ outputs; if the inputs have crystal stability, the beat-frequency output, as measured on a counter, will also have the same stability, even though the scope trace may indicate timing jitter.

Initial checks were carried out using an FT243 crystal on 6.747 MHz and an old General Radio 1001A signal generator. Although it was capable of driving the D input with a necessary minimum 2-volt sine wave, it was found that identical results could be obtained by setting up a 1.3-volt dc bias, and ac coupling a 0.1 V rms or larger rf signal in. This also simplified the buffer circuitry installed later. Setting the generator on 48.229 MHz produced a 1.0-MHz beat with the seventh harmonic of the crystal at 47.229 MHz. This beat was stable to a few Hz for minutes at a time.

circuit adjustment

Satisfied with flip-flop mixing performance, my attention next focused on the MC4044P edge-triggered phase detector which requires TTL logic swings for operation. One input is fine as it comes from the flip-flop. The other input, vfo derived, starts out as a low amplitude sine wave and this must be squared up to develop a proper driving signal. One half of a 7413 dual Schmitt trigger was used initially and worked alright, but in the interest of reducing cost and improving isolation, RTL squaring amplifier was used in the final design.

The jfet source-follower dc output level must be optimized. I found that for this circuit correct adjustment for maximum lock range requires the following procedure: Set vfo to 1.0 MHz. Close push-to-test switch, which applies 1.65 V to the gate, then vary source resistor to obtain either 2.6 V on pin 8 of the 4044 or a mid-scale reading on a 5-mA meter

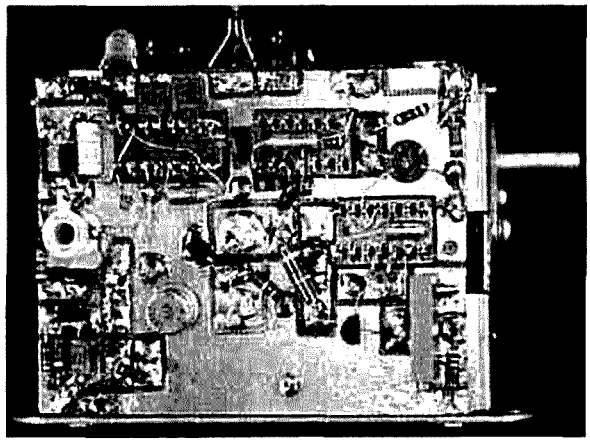
table 1. Tuning and crystal chart for receiver local oscillator use.

direct tracking

receiver frequency (MHz)	vco (MHz)	crystal (use one) (MHz)
50.0/51.0	48.4/49.4	5.32, 5.98, 6.84, 7.98
51.0/52.0	49.4/50.4	5.43, 6.11, 6.98, 8.15
28.0/29.0	26.4/27.4	5.18, 6.47, 8.63
29.0/30.0	27.4/28.4	5.38, 6.72, 8.96

inverted tracking

50.0/51.0	51.6/52.6	5.90, 6.63, 7.58, 8.85
51.0/52.0	52.6/53.6	5.41, 6.01, 6.76, 7.73
28.0/29.0	29.6/30.6	5.18, 6.22, 7.77
29.0/30.0	30.6/31.6	5.35, 6.42, 8.02



Top view of the vfo printed-circuit board with the FT243 crystal in place. Printed circuit layouts are available from ham radio.*

*Complete printed-circuit layouts for the PLL are available from *ham radio* magazine for 25 cents.

(see fig. 1). Next, tune vco to wanted frequency, then release switch.

Phase locking will take hold and may be verified by meter tracking of any further tuning-slug change. Set slug for a mid-scale reading. That's it. Vfo tuning limits will cause a current variation of about plus or minus 1 mA. If a voltmeter is used, pin 8 will be 2.0 V at 1.5 MHz and 3.2 V at 0.5 MHz. Maximum possible swing is 1.0 to 4.5 V.

A pilot lamp and transistor wired as shown in fig. 2 is entirely adequate in lieu of a milliammeter. A steady full *on* or *off* lamp condition when vco tuning is changed means out-of-lock, while half normal brilliance indicates lock-up. Correct circuit adjustment follows the same procedure as before, first setting up 2.6 V on pin 8 of the 4044. With the vfo on 1.0 MHz, alternately depress and release push-to-test switch while varying vco tuning until there is no change in lamp brightness. This corresponds to a midscale current meter reading. Lamp glow will change slightly for full vfo coverage.

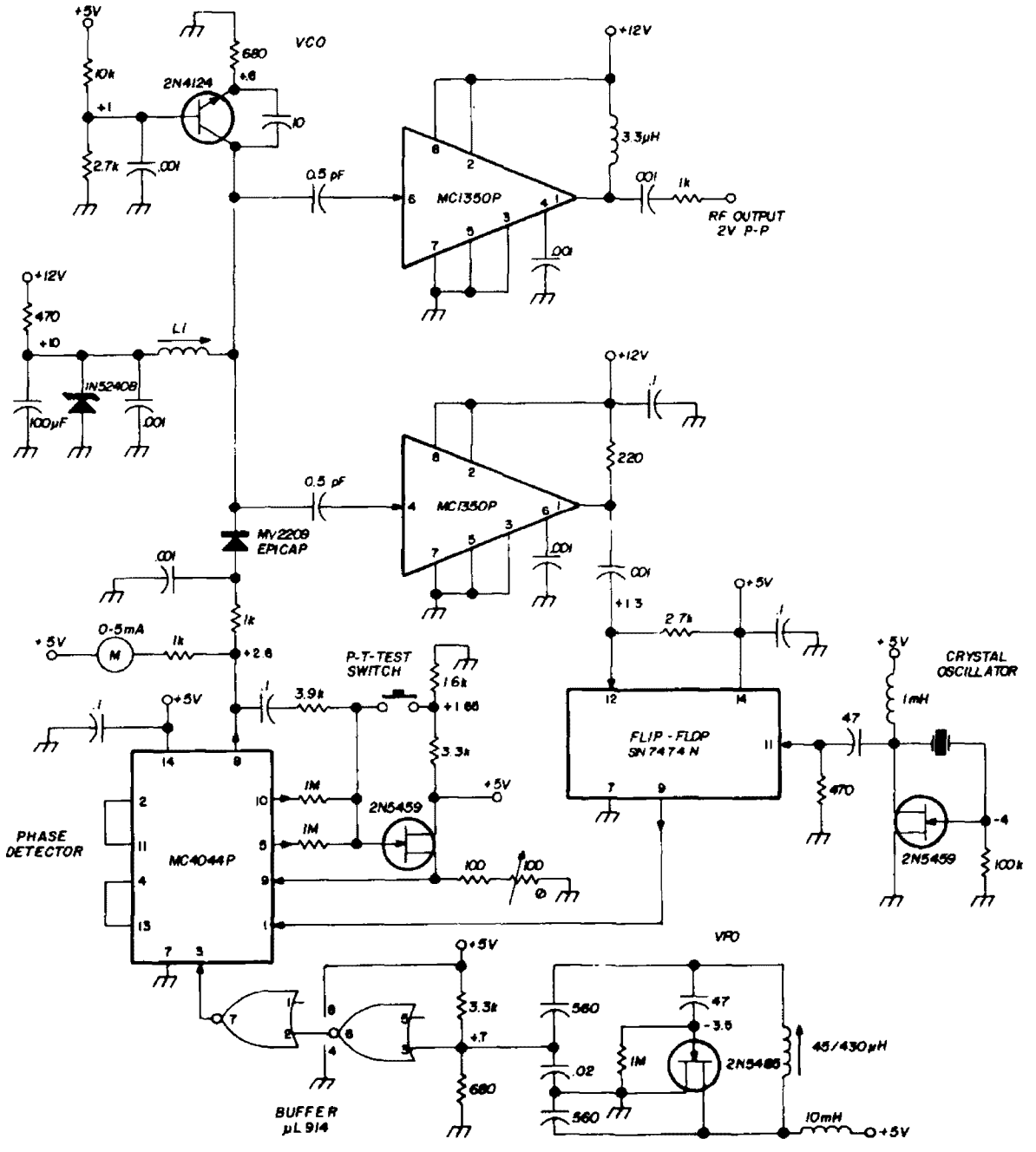


fig. 1. Circuit of the tunabla 6- and 10-meter phase-locked loop. L1 is 6 turns no. 28 enameled (48-54 MHz) on a CTC 3/16" slug-tuned form; use 12 turns no. 28 for 26-32 MHz.

A scope connected to pin 9 of the 7474 is very useful in observing the in-lock and out-of-lock square wave present there. Depending upon vco tuning, its frequency will be anywhere from zero to several MHz unlocked, but exactly the same as the vfo when phase locked. RTL buffer output has unsymmetrical fast rise and fall pulses, with edges being in the order of 30 nano-seconds. A pair of inexpensive MC1350P wideband amplifiers provide excellent isolation and amplification of the vco. This oscillator uses only stray capacitance across the tank coil to maximize pull range of the Epicap diode.

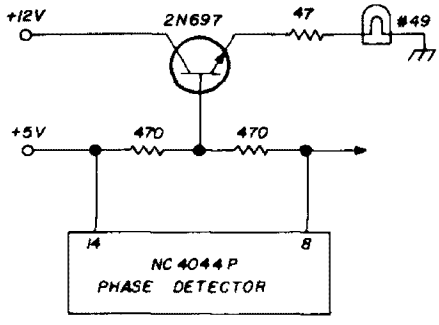


fig. 2. Alternate phase-lock indicator using a number 49 pilot lamp.

The slug-tuned vfo is built from a car radio tuner. Its normal 0.55- to 1.6-MHz spread may be used unaltered, but the high-end signal tends to come through directly on a 1.6-MHz i-f, so this one was fudged to cover .5 to 1.5 MHz. Lacking such an assembly, a Colpitts vfo may be built using a dual-section 420 pF broadcast variable and fixed inductor. RTL buffer drive would be via a link.

An earlier version of this phase-locked loop used a MC4024P dual voltage-controlled multivibrator. One half functioned as a crystal-controlled oscillator while the other half amplified and squared up the vfo signal. Crystal oscillator warmup drift was unacceptable mainly because a high harmonic is used for mixing. Also, chip isolation was insufficient in preventing integral harmonics of the vfo from zero beating with the crystal and developing

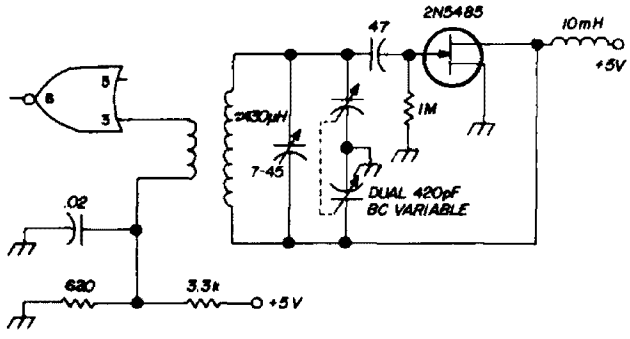


fig. 3. Alternate vfo circuit using a dual-section 420-pF broadcast variable.

enough spurious variation in Epicap control voltage to randomly push the vco frequency around a few hundred Hz. This is because pull range is something like 1 kHz/mV. Therefore, control signal purity is extremely important.

circuit operation

This circuit, in a room temperature environment, has frequency drifts from start-up amounting to about 3 Hz for the crystal oscillator and not more than ten times that for the vfo. Total drift of the locked vco is less than 50 Hz at 48 MHz; a few half-hour runs made during testing checked out at about 38 Hz. For minimum waveform distortion, the untuned buffer should work into a moderate impedance of a few kilohms such as the resistive-loaded input of a dual-gate mixer. The buffer may have a tuned tank and link coupled output if you desire.

Wired as shown, signals at pins 1 and 3 of the phase detector produce direct tracking. That is, as the vfo moves up in frequency, so does the vco. If these leads are interchanged, tracking inverts and this mode may be useful, depending upon

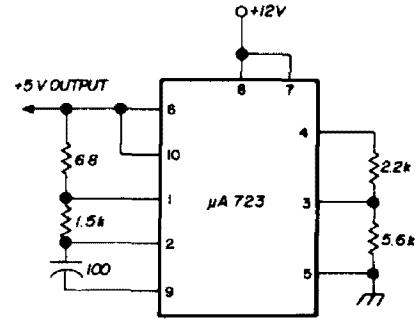


fig. 4. 5.0-volt regulator for the PLL.

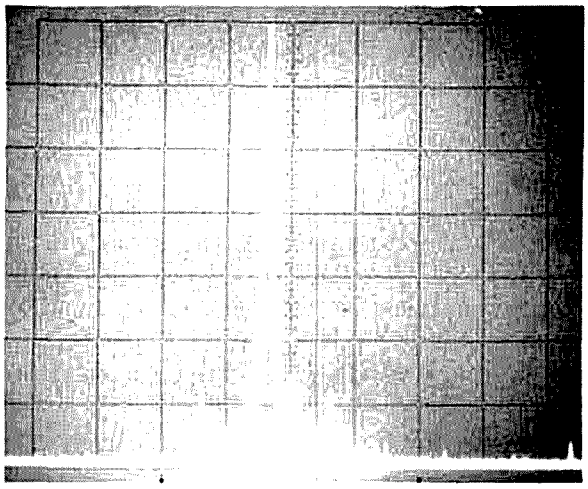


fig. 5. Rf output of the PLL at 50.5 MHz.

what crystals and frequencies are involved. To see how this works, refer to the tuning and crystal chart in table 1 which shown 6- and 10-meter local oscillators in conjunction with a 1.6-MHz i-f and a vfo tuning from 0.5 to 1.5 MHz. There are many possible crystal combinations; only those within the 5- to 9-MHz spectrum have been listed.

Other intermediate frequencies and vfo ranges can be worked out as well as frequencies for transmitter vfo use. It is advisable not to go too low in crystal frequency because of possible mode jumping. Harmonics are less widely spaced and, with the lowpass filter having a time constant of about 1/4 second, any fast transient in the tuning diode network can cause it to shift the vco out of one lock and into an adjacent lock before servo action can pull it back.

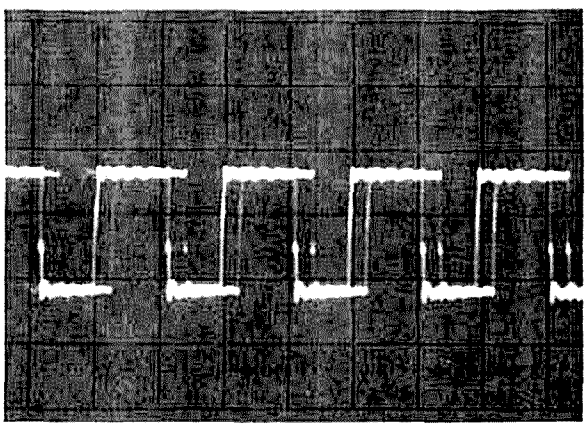


fig. 6. 1 MHz phase-lock signal at pin 8 of the MC4044P phase detector. Signal is 2 volts peak-to-peak.

If test equipment is available to monitor the phase-locked loop, an informative experiment can be run to induce this malfunction by using a low-end 80-meter crystal, vfo set near its high end, vco locked, and rapidly changing the vco slug tuning. These worst-case factors are absent in normal operation, of course, but confidence in the circuit is gained by being aware of its limitations and operating within such bounds.

An additional tradeoff for simplicity is that each 1-MHz tuning range will require another crystal as shown on the chart. Frequencies listed do not have to be matched exactly; vfo dial calibration will take care of it. One exception appears in the transmitter control list (table 2) where a single 5.500-MHz crystal serves for the first 1 MHz on 6 and 10 meters.

table 2. Crystal chart for using the phase-lock signal for transmitter control. The 48-MHz output may be multiplied three times for use on 144 MHz.

vco output MHz	cryatal frequency MHz
50.0/51.0	5.50, 6.18, 7.07, 8.25
51.0/52.0	5.61, 6.31, 7.21, 8.41
48.0/49.0	5.27, 5.93, 6.78, 7.91
48.33/49.33	5.31, 5.97, 6.83, 7.97
28.0/29.0	5.50, 6.87, 9.16
29.0/30.0	5.70, 7.12, 9.50

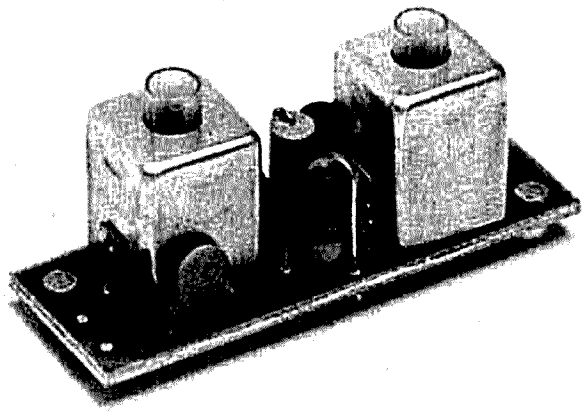
summary

The engineers at Motorola have certainly come up with a very clever circuit that should find its way into many hf and vhf amateur-band receivers because of low cost and high performance. It should also make a dandy low-drift and stable transmitter frequency control. In addition, the PLL technique neatly sidesteps problems of spurious emissions due to unwanted mixer products getting into the output signal..

reference

1. Treadway and Reed, "Generate Stable High-Frequency Signals," *Electronic Design*, 6 January, 1972.

ham radio



improved six-meter preamplifier

A six-meter version of a popular two-meter unit along with hints and improvements for both units

Jerry Vogt, WA2GCF, Rochester, New York 14612

In an article entitled, "Improved Two-meter Preamplifier," which appeared in the March 1972 issue of *ham radio,* I described an easy-to-duplicate, non-neutralized two-meter preamplifier and offered to make kits available to other amateurs. The response to the article was overwhelming. Apparently, there is quite a need for such units; something which comes as no great surprise. Among the replies received were many interesting notes describing experiences of others in trying to build preamps for a variety of uses.

Of course, I did my best to answer the many questions received including the numerous requests for information on adapting the two-meter unit to 50,220 and 432 MHz. The mail prompted not only the design for the new six-meter preamp described here, but also some additional work in refining the existing two-meter preamp. Among other things, the overall size was reduced by half, and

the input circuit was made considerably sharper to provide additional out-of-band rejection. The six-meter preamp in this article reflects the refinements developed for the new two meter unit.

The schematic of the six-meter preamplifier is shown in fig. 1. The design parameters are essentially the same as previously detailed in the March issue. Suffice it to mention that the circuit is an ac-coupled cascode configuration, without the need for neutralization. Therefore, anyone can tweak it to his heart's content, without throwing it into oscillation.

In construction, the six-meter preamplifier is very similar to the one described and illustrated in the previous article. The unit is built on a printed-circuit board, with one shielded coil at each end and the junction field-effect-transistors with their related parts in the center. Coils are a special adaptation of a commercial plastic coil form which is not generally available to amateurs and use 10-32 slugs made of iron 9 material. The preamp may also be made with other coil forms and shields; however, the tricks are to establish a good layout for shielding and to discover how many turns of wire are required. The diligent experimenter with a vhf signal generator and an rf millivoltmeter of some sort should have no great difficulty if he observes normal vhf construction techniques.

For the amateur not so well equipped or the experimenter who doesn't wish to re-invent the wheel, kits and wired units are available, complete with detailed construction information.* Clubs may wish to obtain kits for construction as a group. This type of project has been tried in this area with projects much more involved than this, and they have been very successful. Table 1 lists coil and capacitor data for the preamp. Although you may not duplicate the parts exactly, the table should serve as a guide for selecting alternate components.

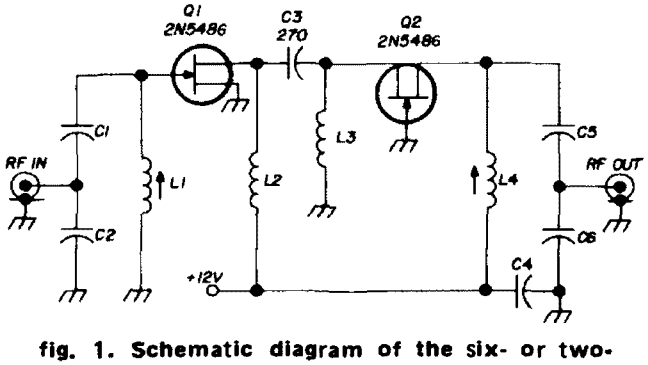


fig. 1. Schematic diagram of the six- or two-meter preamplifier. Parts values are listed in table 1.

table 1. Component values for six- and two-meter preamplifiers. All capacitances are in picofarads.

	50 MHz	145 MHz
C1	20	5
C2	50	20
C3	270	270
C4	680	270
C5	15	5
C6	82	20
L1 (10-32 form)	8.5 t.	3.5 t.
L2,L3	5.6 μH	1.0 μH
L4 (10-32 form)	10.5	4.5 t.

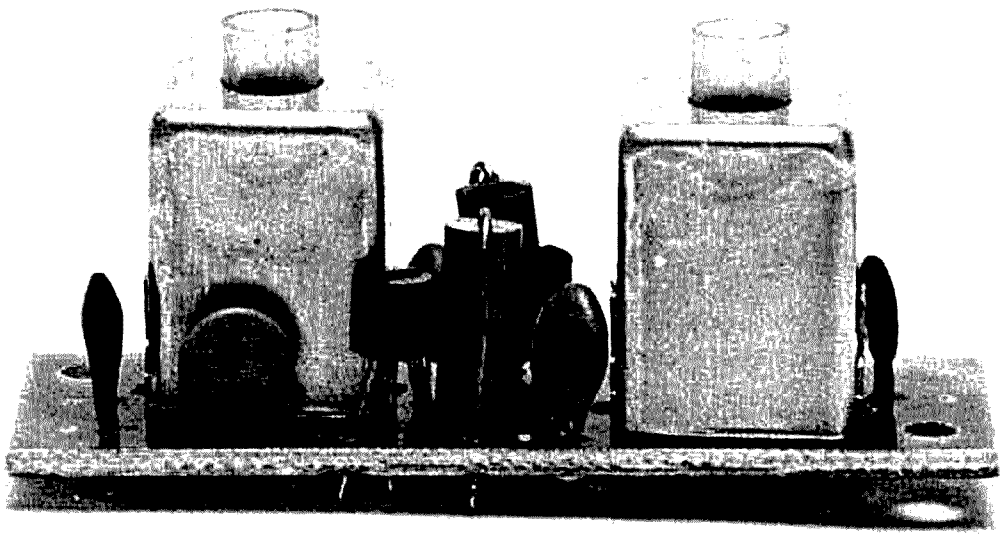
test setup

Several readers have asked how to make a test setup for tuning the preamps after construction. As you might imagine, I use quite an elaborate setup, including an HP-608 signal generator, a specially-developed solderless test fixture with spring-loaded pin contacts and an rf millivoltmeter. The fixture includes a built-in 100-ohm load resistor and a built-in detector circuit although a resistor at the end of your coax cable or soldered to the board will also work. An

*The following are being made available in conjunction with this article. Both 6- and 2-meter models are available; so *specify which you want.* A complete parts kit, including drilled printed circuit board, all components, and complete instructions for construction are available at $6.00, postpaid. Completely built and tested preamps are available at $10.00, postpaid. Special frequencies between 25 and 170 MHz are available at $12.00, postpaid Factory built preamps (only) may be returned for repair for a fixed repair charge of $3 (prepaid) if trouble develops within 90 days (usually damage incurred during installation). Quantity prices are available to clubs and individuals. Contact Hamtronics, 182 Belmont Road, Rochester, New York14612.

rf probe of the type shown in the *ARRL Handbook* will work well for initial coil peaking if you have a very sensitive vtvm (on the order of 100 mV full scale). The important thing is to keep all leads extremely short.

careless in completing the installation in your set. I have had people tell me that they had shorted coax cables, high voltage across the 12 volt input line, coil windings wired to the wrong points in the circuit and other assorted accidents. Tran-



The preamp is built on a small printed-circuit board just over 2-inches long. This photograph is nearly twice actual size.

I use a 100-ohm load instead of a 50-ohm load, since I've found that most receivers present a greater than 50-ohm load to the preamp. Since any high-gain amplifier will take off if you don't load it properly, I do all testing with the high load resistance. There is plenty of reserve gain in the preamp due to the cascode design; so that even if loaded down to 50 ohms or less by the few receivers which do present that low a load, there is ample gain to mask out the noise in the front end of your receiver and improve the sensitivity.

hints

After you have built your preamp, I suggest that you take a good look at some of the hints and kinks in my previous article. You would be surprised at the problems you can generate if you are

sistors are very unforgiving; so check everything before firing it up for the first time, especially in the set. Don't wire it into your transmit rf line (easy to do with a transceiver), don't wire with hook-up wire in lieu of coax (its been done) and don't use a dropping resistor from your 200-volt line without a zener diode (that's been done, too). Look into the power supply section of any good handbook for zener diode regulators if you want to derive your 12 volts from a high voltage line. Handbooks give you all the details on doing it. Make sure that you anchor the preamp down to your set's chassis or case — don't let it rattle around.

I hope that some of these suggestions will help to keep you out of trouble and in good spirits. Have fun with your souped-up receiver. ham radio

zap that transient

How to use transient suppressors to protect sensitive solid-state components from high voltage pulses

Carl C. Drumeller, W5JJ, 5824 N.W. 58 Street, Warr Acres, Oklahoma 73122

No, this is not a dissertation on killing bums or hitchhikers! Rather, it is some advice on preventing transient voltage surges from killing your solid-state components. These surges, which have a duration normally measured in microseconds or, at most, in milliseconds, still can be deadly to devices that have a limited puncture voltage rating. And that takes in most solid-state components!

Transient high-voltage surges come from so many sources and are so easily (universally?) generated that they are inescapable facts of life which one must expect to live with — and to cope with. They can be originated in either dc or ac circuits, and most often get their start in a brief but destructive life by either the closing or the breaking of a circuit having a flow of current. If, for instance, that circuit contains some inductance, breaking the current flow causes a collapse of the lines of force about that inductor, generating a voltage pulse that can range to surprisingly-high values.

How high? Ten times, if there are no limiting conditions present. That means the innocent-appearing 117-volt power line feeding your power supply transformer can be the source of a pulse peaking at 1170 volts! Usually, though, and we can be thankful for this, there are factors present that limit such wild excursions to a little over twice the normal voltage. But don't bet your transistors

and diodes on Lady Luck always riding your shoulder; she's notoriously fickle! Those fortutious circumstances that save our necks (and diodes) usually take the form of intentional or accidental capacitances, resistances and *inductances* that serve to soften the abrupt thrust of the escalating voltage. A transformer is an excellent example. Fortunately, it's less expensive to produce a power transformer with a poor frequency response; therefore, our 60-Hz transformers simply balk at passing those high-frequency components of the steep wavefront of a transient pulse. Then, too, the effects of turn-to-turn and layer-to-layer capacitance add soothing and smoothing.

If, however, you're using a rectifier connected directly to the power line (a dangerous practice itself), there's no transformer to protect your diode; so you'll have to rely on other factors such as the cold resistance of a 117-volt pilot lamp across the line. There are many circuits, usually involving capacitors and resistors, that help to swamp out the undesired surges.

transient suppressors

To do the job right, though, you should consider some of the devices marketed especially for transient suppression. The trademarks vary with the manufacturers, of course, but these suppressors can be classified into three catagories: spark gaps, capacitors and non-linear resistors (including semiconductors). Let's ignore the first two and concentrate on the last one. In this category, you're looking for a resistor that, for the voltage normal to the device you want to protect, offers a very high resistance. For voltages somewhat higher than this normal value, though, you'd like for the resistor to present a very low resistance and to be able to handle (for a few milliseconds) a very high current.

Sounds like cloud nine dreaming, doesn't it? Yet, there are quite a number of such varistors available. They may be made of silicon carbide, they may be silicon zener diodes, they may be selenium diodes or they may even be gas discharge tubes. Other types exist, but manufacturers do not like to reveal their exact compositions.

An ordinary resistor obeys Ohm's law: $E = IR$. A varistor, however, follows a variation which looks like this: $I = KE^{\alpha}$,

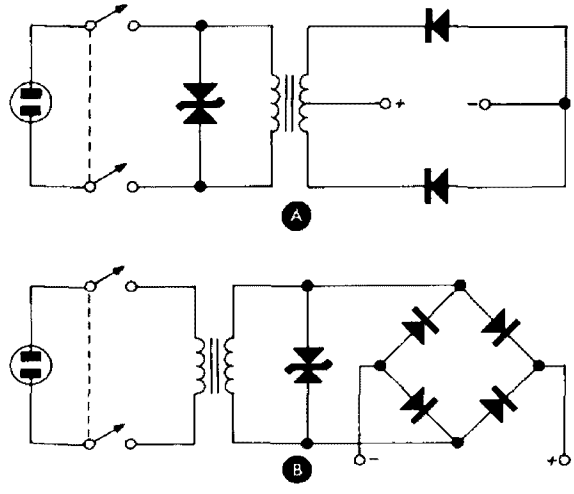


fig. 1. Using transient suppressors for effective protection of solid-state components. In (A) varistor is installed in the primary of the power-supply transformer. In (B) the varistor is connected across the secondary winding.

where K is a constant dependent upon the composition and size, and α is what you might call the figure of merit for its performance as a varistor. This is unity for a normal resistor, of course, and may be as high as 70 for some of the better designs. Even the run-of-the-mill varistors have an alpha of five or better. So, as you can see, the current that varistor will drain off really shoots up as you raise the voltage.

Most, but not all, varistors have a knee in their I-E relationship. Below this knee, the varistor may behave more or less like a normal resistor. Above the knee the alpha factor becomes evident. You'd like, of course, to have a varistor with a sharp knee just above the highest voltage you expect to encounter in normal operation of the circuit to be protected. By consulting the curves provided by the several manufacturers of varistors, you can select

one with suitable characteristics. These curves need to be read with care, though, as they can give illusionary information when either the ordinate or the abscissa scale is depicted in other than the units directly applicable to the intended use. Curves plotted on log-log graphs are the least likely to be misleading.

commercial transient suppressors

Several manufacturers produce varistors, each identified by a registered tradename. General Semiconductor Industries Inc., calls their devices Trans-Zorbs. These are silicon semiconductors, and, although their mode of operation is not given in the literature, I suspect that they're zener diodes. Packaged in either metal or plastic, they look like small diodes. For up to 1/120th of a second, which is much longer than the duration of a transient, they'll handle 200 amperes. You can buy them for protecting circuits carrying dc voltages from 5 to 190 volts. For a 5-volt logic power supply, you'd want the ICTE-5, which sells for $4.50 in single quantity. For 15-volt ICs, the 1.5KE16A, at $2.30, would do the job. The 24-volt 1.5KE27 is the same price. For protecting the input side of a power transformer, the 1.5KE150, costing $2.50, would serve. These can be purchased from General Semiconductor at 230 West Fifth Street, Tempe, Arizona 85281.

International Rectifier uses the trade mark *Klip-Sel* for their transient killers. As you might suspect from the name, they're selenium cells of special design, and are available in either polarized (for dc circuits) or nonpolarized (for ac circuits). Allied Electronics, 100 North Western Avenue, Chicago, Illinois 60680, is one of the firms that will sell *Klip-Sels* in single quantity. Because of the rather high minimum voltage at which a selenium cell starts conduction, *Klip-Sels* are best adapted for use in the primary, 117-volt side of power supplies. A polarized *Klip-Sel,* KSA6DPF, suitable for such an application, costs $6.25. For moderate power, the KYP6DPF at 83 cents will do the job. This is pretty cheap diode insurance.

The varistor most recently marketed is the *GE-MOV*, which is of the metal oxide type, combining both high alpha and small physical dimensions. Because of the range of voltages in which it is available, 150 to 1000, it is suitable for use in the primary circuit of stepdown transformers or either the primary or secondary circuits, within voltage limitations, of step-up transformers. The *GE-MOV* will handle extremely high short-duration current peaks, well above 1000 amperes. This capability is coupled with quite low costs. The VP130A10, which is adapted for use in 117-volt circuits, is priced at $1.80 in single-lot quantities. It may be purchased from any of the General Electric distributors, which are located in most large cities. Newark Electronic, 500 North Pulaski Road, Chicago, Illinois 60626, is a centrally located source.

All of these varistors have one application characteristic in common: they are selected not to start conduction immediately upon a voltage excursion just above normal, but to exert their full clamping effect at approximately 2.5 times the applied rms voltage. Thus, there is no appreciable power dissipation in the varistor under normal operating conditions or for very mild transients. Of course, you recognize the presence of these mild transients; that's why you normally specify and use solid-state rectifiers that are rated at several times the expected voltage.

Now that you know just about all you need to know about a varistor, there might remain one question: Where do you use them? The accompanying schematic diagrams show the answer. There's one problem in drawing these diagrams; each manufacturer uses a different symbol to identify his product. I've selected one, but this selection by no means is to be taken as a recommendation for that manufacturer's product over the others. They're all good, all effective, all worthy of being in every power supply.

ham radio

solving overload problems with vhf converters

Straightforward ideas for locating and fixing pesky overload problems in high-performance vhf converters

Hank Cross, W1OOP, 111 Birds Hill Avenue, Needham, Massachusetts 02192

For many hams, especially those out in the wide open spaces, the main problem in vhf receiving equipment is to hear *something.* The average ham's thinking on the subject of sensitivity and noise figure is a lot clearer than it was twenty years ago, and sensitive receiving equipment is standard these days. However, when the vhf bands open and the ham across town comes out from under his rock and fires up his two-meter kilowatt, another problem shows up: overload. It's serious in New England when those guys are slugging it out from the mountaintops at contest time, and a major problem in North Jersey about any time the band is good. If we had a clear idea of what's causing the trouble maybe we could fix it. Let's try to sort out the causes of the problem.

One is the other guy's lousy transmitter. To prove that the trouble is from this cause, you need to show that the strength of his signal is *not* the important factor. An attenuator, or simply turning the beam so that he's knocked down to a reasonable level, will check this factor. If, at a mild S9 level, he still has splatter and buckshot all over the next two hundred kHz, maybe you should call him on the telephone and complain. Usually, attenuating his signal fixes things pretty well, and that implies that you have work to do on your receiving apparatus, even if he *is* overmodulating.

Which brings us to overload in the converter rf stage. My own experience is that less rf gain is better, but that the particular kind of device you're using in the front end (tube, fet or bipolar transistor) has only a minor effect on the problems most vhf amateurs have — which kind only matters when other effects are reduced by ten or a hundred times. I *do* know that the problem can be licked, i.e., we can make low-noise preamps which will show no trouble from

strong signals until the stages following have completely overloaded.

the mixer

As any sideband type can tell you, an amplifier which puts out about ten watts as a linear amplifier will flat-top at around a watt when used as a mixer. Transistors do this, too, and that means that the mixer overloads before the rf stage. Diode mixers or fet mixers will handle more input signal; they also take more local-oscillator drive. Anzac lists a double-balanced mixer (MHSM-3) which is linear within one dB up to 30 milliwatts, provided you have 200 milliwatts of local-oscillator power. (Conversion loss is 9 dB.) The first i-f amplifier, of course, must be linear with four milliwatts of input signal! The stage after *that* is beyond the scope of this discussion.

If you use a crystal-controlled converter which passes a relatively wideband (if all the QRM is within a 200-kHz wide band, *that's* relatively wide) the receiver following is possibly a source of trouble. In my experience, certain military receivers are good for use following a converter, but some amateur receivers are subject to overload troubles, even though they seem excellent when used barefoot on the lower bands. The R390, BC348 (with the original amount of rf gain), BC342 and RBC are particularly good. (The only amateur communications receiver I can say that about is the original Collins 75A.)

Intermediate frequency (the converter's output frequency) also has a bearing on the results: most receivers have better rf selectivity at, say, forty meters than on the ten-meter band, so the usual receiver following a c-c converter will have overload trouble plus or minus 50 kHz at 7 MHz, say, and ±100 kHz at 14 MHz and 200 kHz or more on 28 MHz. Some receivers used in combination with a high-gain converter should have a pad (6-dB for instance) between the converter output and the receiver input. A switchable pad (Kay or Waters) installed between converter and receiver is a good

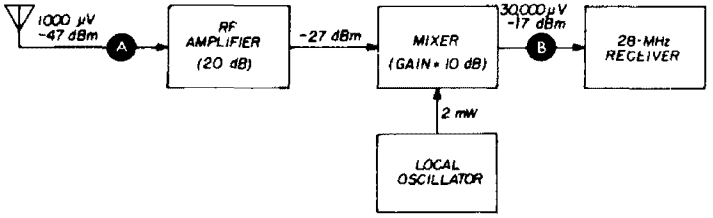


fig. 1. High-performance vhf receiving setup. The attenuating pad at A checks the other guy's transmitter. The pad at B determines whether the overload problem is in the converter or the receiver (see text).

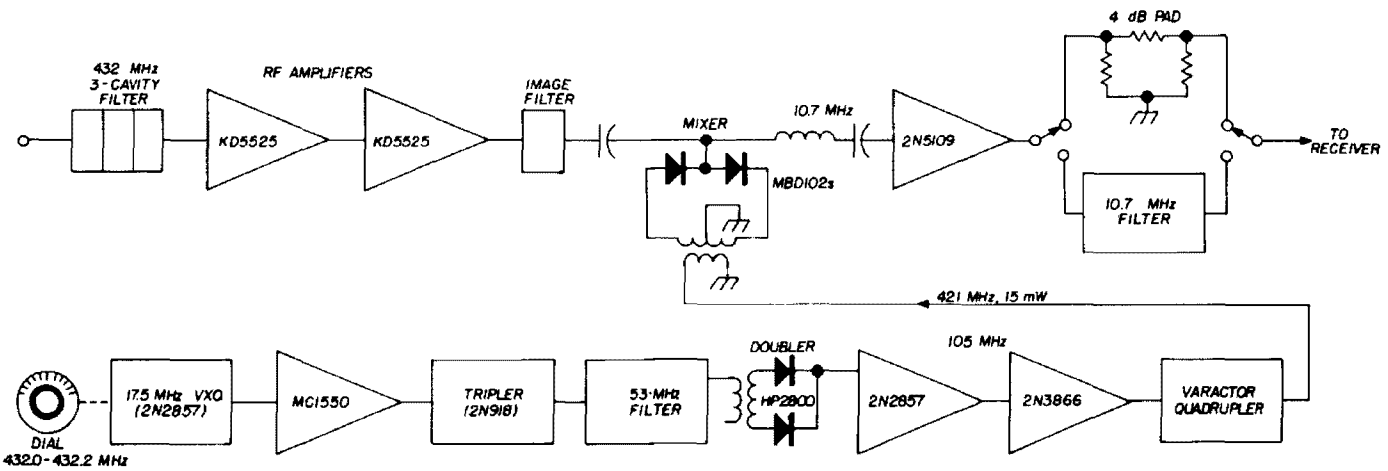


fig. 2. The 432-MHz vxo converter used by W100P. The MC1550 IC is used as an amplifier-limiter to level the 17.5-MHz output from the vxo.

thing to try. If attenuation helps, there must be an optimum value.

For the special case of a band where all the activity is within iess than 200 kHz, there is an excellent solution: Vary the frequency of the first oscillator, and put a crystal filter in as early as practical at the first i-f. At 432 MHz I used a vxo which had up to 400 kHz of range. With about 20 dB of rf gain, it seemed reasonable to use a crystal mixer followed by a low-noise high dynamic range first i-f amplifier which in turn fed a 10.7 MHz filter (four crystals) before the second conversion. The crystal filter was arranged so it could be switched in and out of the signal path, (with a 4-dB pad replacing it to keep levels the same). It was easy to tell if it was of any benefit. It was. See fig. 2.

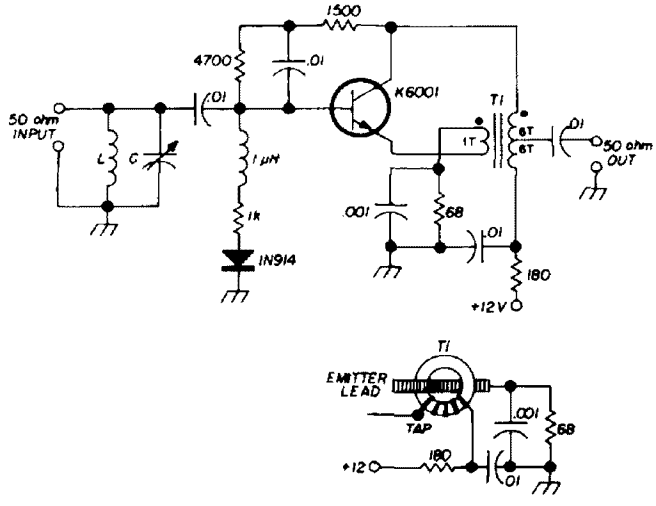


fig. 3. Vhf feedback amplifier for 144 MHz. Without the tuned circuit at the input performance is only 3 dB down on 50 and 220 MHz. Gain is about 15 dB on two meters, noise figure is 1.5 dB at 15 mA. Transformer T1 is 6 bifilar turns on Ferroxcube 1041CT 3E2A core. Keep the emitter lead as short as possible.

amplifiers and overload

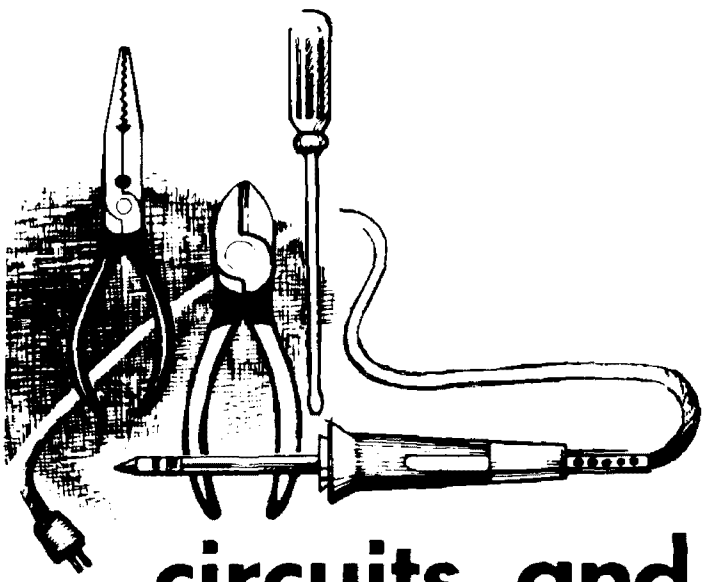
Let's think about an amplifier. It has some value of gain. It has a noise figure. And it has, for one way to specify it, a "one-dB-compression point." If we assume that the amplifier is exactly linear up to, say, ten milliwatts output, and that it limits sharply at that point (flat-tops) then as we run the input signal up to the limiting level and 1 dB beyond, the gain will be 1 dB less than for a signal low enough to stay within the linear range, i.e., it is compressing by 1 dB.

For a particular value of gain, there is obviously a miximum value of signal that can be amplified linearly. The *output* compression point, for a vhf amplifier, depends mostly on current and load resistance. For low-level amplifiers, the bias point is usually optimized for noise figure.

If a survey is made, you will find that optimum NF for most vhf transistors occurs at currents between 1 and 5 mA. Among these, those which exhibit a given NF at higher current are preferable. Some types, which have optimum NF at 1 to 2 mA, may still be very quiet at 5 to 15 mA. One commercial transistor (2N5109) has good NF from 1 to 15 mA, best usually at 3 mA, but specified as 3 dB max at 200 MHz at 10 mA. In terms of signal-handling capability and noise figure it compares favorably with a fet.

Somewhat higher priced types (MS-175, K6001) will give 1.5 dB NF at 15 mA at 150 MHz. In a feedback amplifier gain is 15 dB and the output compression point is over 20 milliwatts (see fig. 3). At 2 mW out, that is, two two-milliwatt signals, the in-band intermodulation product (third-order product) is more than 40 dB down. Such an amplifier is still operating with good linearity when succeeding stages are overloading, and therefore, there is little point in worrying about how to further improve the *first* stage.

ham radio



circuits and techniques

ed noll, W3FQJ

antenna tuning units

Antenna tuners come in a variety of different sizes, circuits and tuning arrangements. Most are homemade, but some commercial units are available, including the low-power Ten-Tec AC5, and the higher power Johnson Matchboxes and Drake Matching Networks.

Antenna tuning units are used primarily for transmission-line impedance conversion (high to low or low to high), but they may also be designed for converting from one type of feedline to another (balanced to unbalanced or vice versa).

Three very common applications for antenna tuning units (or ATUs) are shown in **fig. 1.** The impedance at the input end of the coaxial transmission line is matched to the output impedance or the transmitter. This may be an unbalance-to-unbalance match such as shown in figs. **1A** or **1B,** or an unbalance-to-balance match as illustrated in **fig. 1C.**

The antenna system impedance can be several times higher (or several times lower) than the transmitter impedance, and a good match can still be obtained with a relatively simple circuit such as the L-network, pi-network or T-network shown in **fig. 2.** I personally prefer the T-network.

The L-network in fig. **1B** is the most common antenna tuner configuration for matching random-length, single-wire antennas. The multiple tapped coil accomodates the random wire length over a wide span of operating frequencies. The variable capacitor sets resonance and influences the impedance ratio needed for satisfactory matching.

The arrangement of fig. **1C** links a balanced antenna and transmission-line system to the low-impedance and unbalanced output of a transmitter. Resonant tuning is provided by the split-stator variable capacitor, while impedance matching is mainly a function of turns ratio and coupling between the separate coils.

atu circuits

A simplified circuit for the Johnson kilowatt matchbox is shown in **fig. 3.** An untuned link, **L1,** transfers transmitter

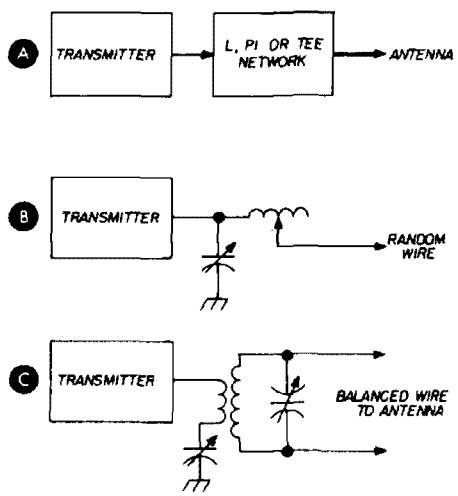


fig. 1. Common applications for antenna tuning units.

output power to a multiturn and band-tapped secondary coil, L2. Resonant tuning is the responsibility of variable capacitor C1. Loading and matching are handled by the dual-differential capacitor C2.

The Johnson Kilowatt Matchbox matches balanced antenna system inputs between 50 and 1200 ohms, and unbalanced, between 50 and 200 ohms. This is not to say that the tuner cannot be used to match low impedance antennas (impedances lower than 50 ohms). If you want to match a very low impedance antenna, such as a beam, without an antenna matching section, you can still obtain proper matching to the transmitter by using a transmission line of a proper overall length. A very low impedance antenna can be made to reflect an impedance higher than the characteristic impedance of the transmission line by using an overall transmission-line length that is some odd multiple of a quarter wavelength, fig. 4.

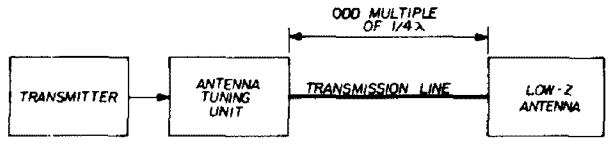


Fig. 4. Tuning a transmission line attached to a low-impedance antenna.

Keep in mind that a mismatch at the antenna can be reflected to the receiving end of the transmission line as an impedance higher or lower than the surge impedance of the line by regulating the overall length of that line.

balance-to-unbalance versatility

If leads are brought out from the various components of an antenna tuner, it can be made to have added versatility as shown in **fig. 5.** This arrangement also includes a tuned primary for further optimizing performance and obtaining an exact impedance match. In this circuit the impedance of the parallel resonant circuit can be made high, medium or low, depending upon how the terminals are interconnected.

A low-C parallel circuit is obtained by connecting series-connected capacitors in parallel with the coil. This can be done by joining 1C and 2A. The transmission line is attached by joining 1A to 1B and 2C to 2B. A high-C matching resonant circuit is obtained when 1C is linked to 2C and 2A to 1A. Output is again provided by connecting 1A to 1B and 2C to 2B.

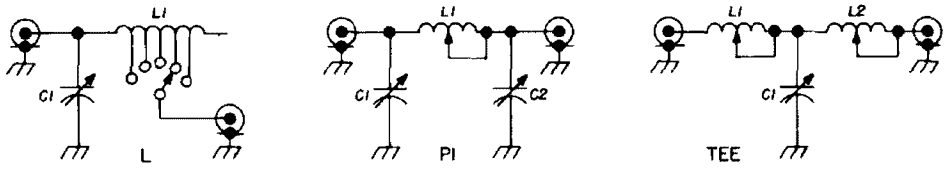


fig. 2. Three basic antenna tuner networks, the L, the Pi and the T.

In low-impedance series tuning, the capacitors must be connected in series with the coil. To do this you need only connect 1B to 1C and 2B to 2A.

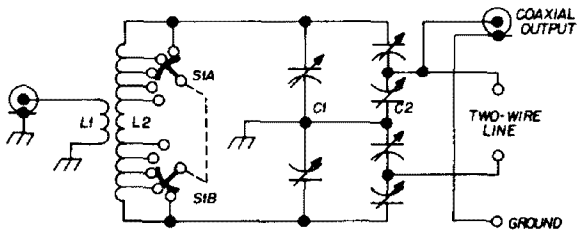


fig. 3. Partial schematic diagram of the Johnson Kilowatt Matchbox.

unbalanced-to-unbalanced T-tuner

The T-network is an excellent tuner for an unbalanced system. It has great range and versatility and can be used to match almost anything connected to the antenna end of a coaxial feed line. It will also function in the same manner when matching a random length of antenna wire as well as a Windom antenna, fig. 6.

The T-section antenna tuner is basically a low-pass filter consisting of two

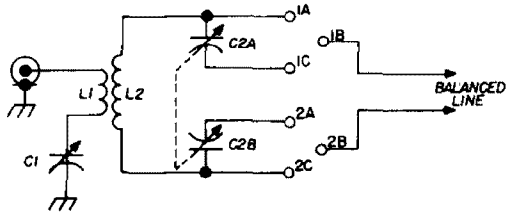


fig. 5. Versatile antenna tuning circuit for balanced transmission lines.

series-connected inductors along with a capacitor connected between their junction and ground as shown in fig. 2. Although there is some interaction between the two coil sections, L1 and L2, the value of L1 has a significant influence on the matching between the antenna system and the tuner, while inductor L2 has a greater influence on matching between the tuner and the transmitter. Interaction between the two coil sections can be minimized by mounting them at right angles to one another. Capacitor C1 establishes the proper resonant condition and, if adjustable, acts as a fine tuning adjustment.

For a specific case of matching on a particular frequency, the basic equation of the T-network is:

$$X_{C1} = \sqrt{Z_{IN} R_T}$$

where

Z_{IN} = the input impedance to the line, and

R_T = the output impedance of the transmitter.

Multifrequency and multi-band operation requires the inductors be tapped and the capacitor made variable. Two seven-position switches permit operation on bands 1.8 to 54 MHz. Each coil consists of 30 turns of number 14 wire, 2-5/8-inch length with a diameter of 1-3/4 inches (this is similar to Air Dux 1411 coil stock).

Some experimental work with tap positions will permit optimum performance on each band. In my case ten-position switches were used and taps were placed on the coil so as to decrease distances between taps toward the low inductance end, fig. 7. This arrangement permits greater versatility, and I have yet to connect an antenna that could not be made to load the transmitter.

A 50-pF variable is used for 6 through 20 meters. When operating on 40, 80 and 160 meters an additional two-gang 365-pF variable (sections connected in parallel) is switched into the circuit.

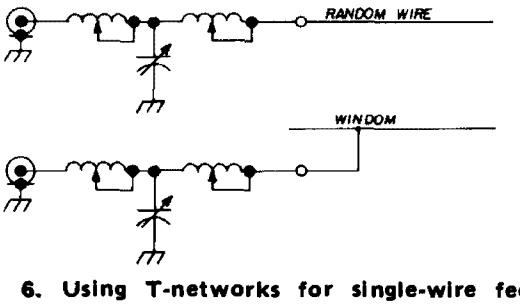


fig. 6. Using T-networks for single-wire fed antennas.

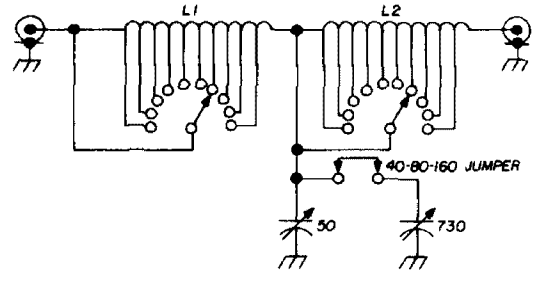


fig. 7. Versatile T-network antenna tuner. L1 and L2 are each 30 turns, tapped at 2, 4, 6, 8, 10, 13, 17, 21, 25 and 30 turns.

When adjusting the tuner you will soon learn the switch positions that favor each band, and, as expected, less and less inductance is needed, the higher the frequency band. However, ideal matching requires some experimentation with each antenna type to find the two most favorable switch positions. Switch positions are found that result in a very low, minimum swr. As the switch positions are selected the variable capacitor is tuned for minimum swr.

You should keep a log for any given antenna so the tuner settings can be quickly changed when you make a band change. If another antenna is used, optimum settings are not likely to be the same. This is a tuner that can be used to load anything, but it does require some initial pre-adjustment to locate the ideal settings for any given antenna. Remember that the tuner can be made to load

anything but this does not guarantee that the anything you use will function as a good antenna.

T-matching at antenna

In commercial radio services the T-match is popular when the matching is done at the antenna. Three common configurations are shown in fig. 8. If the antenna is inductive, the inductance is tuned out by using an input capacitor, C1, which has the same reactance. The T-network must then only match a resistive component to the transmission

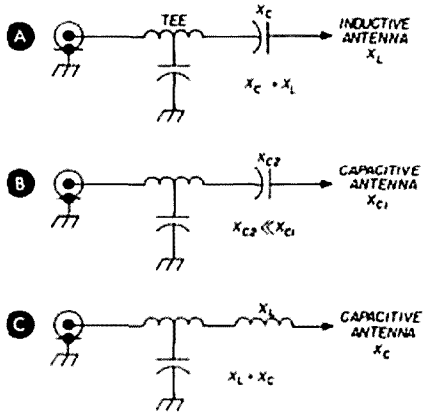


fig. 8. T-networks for impedance matching at the antenna.

line. If the antenna is capacitive, the capacitive reactive component must be tuned out by the input coil of the T-network. An alternative plan, shown in C, uses an input coil, L1, which has the same reactance as the capacitive component of the antenna.

In multifrequency and multi-band operation, as in amateur practice, the components must be variable. The T-network tuner of fig. 7 is ideal for this type of operation. It performs particularly well when vertical antennas are to be matched at the antenna, and the swr on the transmission line between the tuner and transmitter must be reduced to an insignificant value.

Antenna tuners are a great addition to the amateur station and should be considered essential devices for every ham antenna experimenter.

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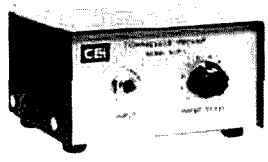
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noise bridge for impedance measurements

Novel modifications to an existing circuit allow measurement of reactive as well as resistive components of unknown impedances

G. Pappot, YA1GJM, Kabul, Afghanistan

A useful instrument in the station of the active amateur is one that will measure resistive and reactive components of an unknown impedance throughout the hf range. Such an instrument is described here. The idea for its construction was brought about by a previous article in *ham radio.*[1]

features

Using a communications receiver as a detector, this instrument will measure:

1. Antenna impedance.

2. Electrical length, velocity factor, and characteristic impedance of coaxial transmission lines.

3. Input impedance of rf amplifier circuits, small inductors, capacitors, baluns, and other rf transformers.

The instrument is effective over a range of approximately 2-30 MHz. Construction is not difficult and readily available parts are used. Power is supplied by a single 9-volt transistor battery. The dials for the reference components are direct reading.

circuit

A diode noise generator, 3-stage amplifier, and rf bridge comprise the circuit (fig. 1). The amplified noise-generator output is coupled to a small toroidal rf transformer. A receiver connected to the DET terminal will indicate a noise null (S-meter or receiver audio output) when the impedances at each end of the transformer secondary are equal, indicating bridge balance. Calibrated scales are used for the controls of components C_x, R_x to measure an unknown impedance, Z_x. The unknown impedance can be measured in terms of equivalent parallel C_x, R_x within the following limits:

R_x: 0 to 250 ohms (resistive)
C_x: 0 to +70 pF (capacitive)
0 to -70 pF (inductive)

If, when measuring Z_x, a negative

value is shown on the dial for C_x, the unknown will have an inductive reactance component, where the inductance can be found from

$$L_x = \frac{-1}{\omega^2 C_x}$$

where

L_x = unknown inductance (H)
$\omega = 2\pi F$ (Hz)
C_x = reference capacitance (F)

of 8 turns.* In the schematic (**fig. 1**), the large dots indicate the beginning of each winding.

Control R_x has a scale that reads between zero and 250 ohms, and control C_x has a linear scale reading between -70 and +70 pF. The scales can be calibrated using resistors and capacitors of known accuracy connected in parallel with the Z_x terminal. Two-watt composition resistors in the 10 to 150-ohm range have

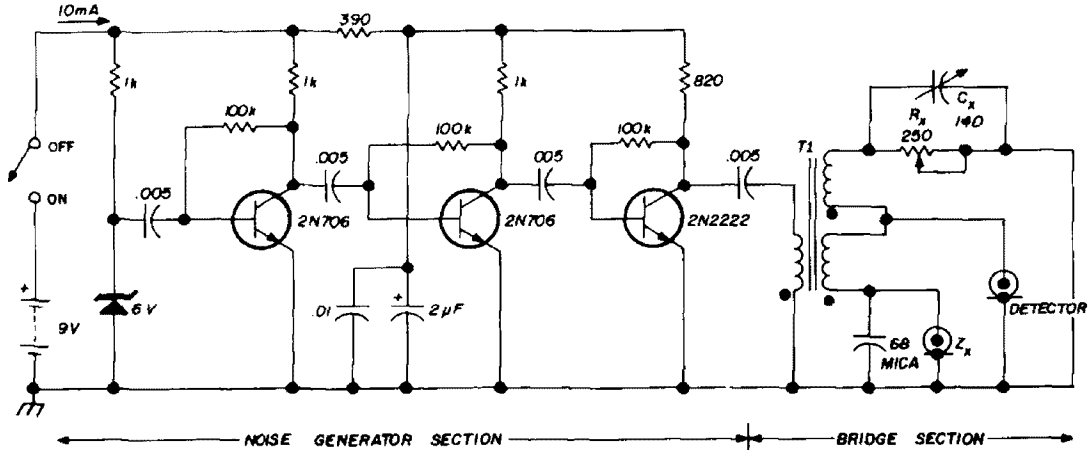


fig. 1. impedance-bridge schematic. T1 is trifilar wound on a toroid core (see text). Reference elements C_x, R_x are adjusted to produce a null when the impedance to be measured is connected to the Z_x terminal. A sensitive receiver is connected to the DET terminal.

More practically,

$$L_x = \frac{-25{,}300}{F^2 C_x}$$

where

L_x = unknown inductance (μH)
F = frequency (MHz)
C_x = reference capacitance (pF)

construction

The entire circuit including battery is mounted inside an aluminum box measuring 4 x 7 x 10 cm (1.5 x 2.75 x 4 inches). The noise generator section is mounted on a PC board. Capacitor C_x is an air dielectric variable with *linear* response. Resistor R_x is a 250-ohm carbon composition pot, CRL No. ACS9-251-U. Terminals for Z_x and DET are general-purpose vhf connectors, type SO-239.

The bridge section should have leads as short as possible. The toroid core for the bridge transformer has an O. D. of 9 mm (0.35 in.). Windings are trifilar, consisting

approximately -14 pF (inductive) reactance in parallel with their indicated resistance value. Thus a capacitor of 14 pF must be connected in parallel with such resistors to cancel the inductive reactance and produce a pure resistance for use within the frequency range of the bridge.

operation

The unknown impedance is connected to the Z_x terminal. If antenna impedance is being measured, care must be taken to use a transmission line whose length is a multiple of one-half wavelength.

Connect a receiver to the DET terminal and switch on the noise bridge. The receiver S-meter will indicate the noise input. Adjust the receiver to the fre-

*A suggested toroid core is the Micro-Metals T-37-10 available from Amidon Associates, 12033 Otsego St., North Hollywood, California 91607. See Hank Olson's article in *ham radio*, April, 1971, page 21. editor.

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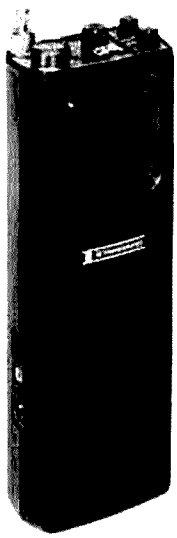
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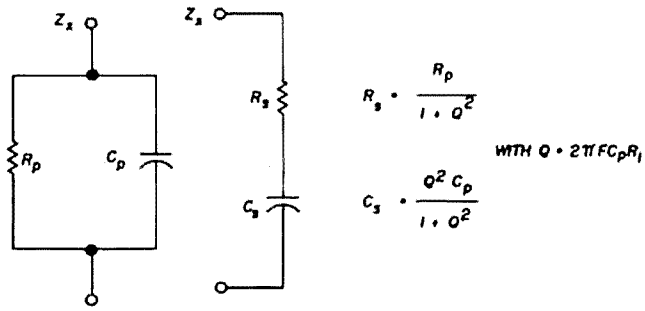
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quency at which Z_x is to be measured. The dials of R_x and C_x are now adjusted for a noise null. When the bridge is balanced at maximum detector sensitivity, the R_x and C_x dials will indicate the parameters of the unknown in terms of equivalent parallel components of resistance and positive or negative capacitance. The dials are direct reading and independent of frequency.

The following expressions can be used to find Z_x in terms of equivalent series-connected components:



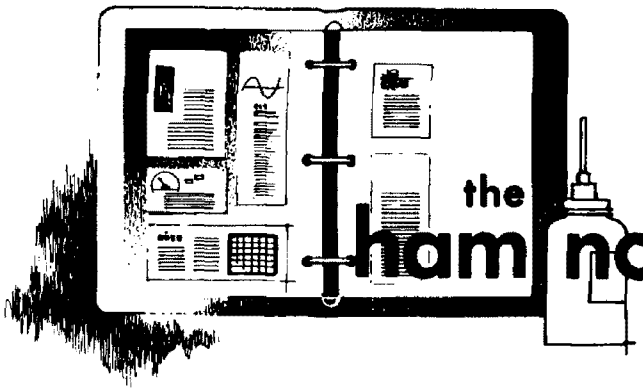
detector notes

I have used a Collins R390/URR and a KWM-2A as detectors in this circuit. With both receivers connected in parallel with the DET terminal, the R390 produced a very sharp null while that from the KWM-2A was relatively broad. This effect was noticed only when measuring frequency-dependent elements (antennas, tuned circuits, etc.). The effect did not occur when measuring pure resistances. Apparently the bridge produces a noise-spectrum gap exactly at the measurement frequency. This wideband gap, as presented to the receivers, provides an opportunity to measure receiver performance with respect to spurious signals much in the same manner that telephone wideband amplifiers are tested using whitenoise generators.

references

1. Don Nelson, WB2EGZ, "The RF Bridge," *ham radio*, December, 1970, pp. 18-21.
2. Hank Olson, "How to Use Ferrite and Powdered-Iron for Inductors," *ham radio*, April, 1971, pp. 15-27.

ham radio



the ham notebook

eliminating tuner overload

I have been building and using crystal-controlled converters with various 80-meter receivers for a number of years. To keep panel controls as simple as possible, all of these converters were built without rf gain controls. Consequently, the rf gain on each tuner used had to be juggled to prevent the converter from overloading the front-end of the tuner. Using my latest solid-state tuner, the problem became particularly annoying. While the obvious solution was to add an rf gain control to each converter, this would mean adding a negative bias supply as well, and the whole thing would become cramped and messy.

I considered the idea of inserting an attenuation pad between converter and receiver and came up with a simple method which works well and takes five minutes to install.

Each of the converters is link-coupled to the output connector. The easiest method of installing a pad is to merely insert one resistor of the correct value in series with this output link. Depending on the tuner being used, this resistor may be as much as five kilohms. With a 50-ohm input impedance receiver, this would represent a 20-dB attenuation between converter and receiver. In my case, I first tried a 2700 ohm resistor which worked so well I did not experiment further.

While inserting this resistor means the converter no longer has 50-ohm output impedance, it is doubtful that this will create any problems unless the tuner used is particularly unstable. The benefits gained in not having to juggle gain controls from band to band will be immediately obvious.

Mike Goldstein, VE3GFN

hot etching

I have found a successful method to keep etching solution warm during a printed circuit project. The only equipment necessary is the plastic container in which the solution is stored, a wash basin and a modified bleach bottle (see fig. 1).

Heat the solution in its plastic container "baby bottle" fashion. When the water is too hot to put your finger in, turn off the heat. The solution is now ready. Pour the solution into the bleach bottle along with the board to be etched. Place the bleach bottle in the wash basin filled with hot water so that it floats. The hot water in the basin will keep the solution warm throughout the process. If agitation is needed, position the basin under the hot water spigot so water can flow into it creating a rocking motion. To prevent the bottle from rotating, scotch tape two pieces of string from the sides of the basin to the opening in the bleach bottle. In a few minutes, the board is etched.

Of course, the bleach bottle can be replaced by any suitable tray. However,

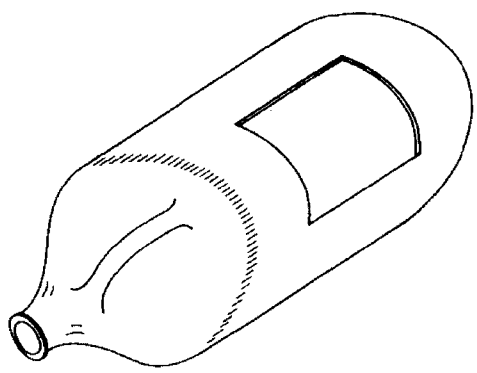


fig. 1. Using old bleach bottle for etching printed-circuit boards.

the big advantage comes when the etching job is done. Merely unscrew the cap and pour the solution back into its container without spillage as with flat trays.

Joseph Turkal, K8EKG

regulated 5-volt supply

The power supply shown in fig. 2 is useful for powering your latest TTL IC project. The unit features easy construction, *low noise* and *good regulation.*

The MOV on the schematic is a General Electric metal-oxide varistor — which performs similar to back-to-back zener diodes. It provides line transient protection. I used a VP130A10, which sells for $1.80. Voltage regulation is by a Fairchild μA7805, three-terminal voltage regulator. Be certain to mount this device with a good heat sink if you want to draw the full rated current. The regulator chip sells for $2.20. Do not eliminate the bypass capacitors C1 and C2 — they serve a very useful function.

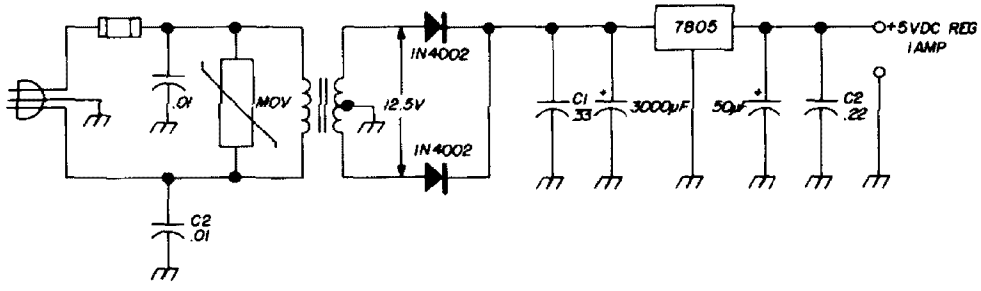


fig. 2. Simple 5-volt regulated power supply.

The regulator and MOV are available from Hamilton Electro Sales, 10912 Washington Boulevard, Los Angeles, California.

Hilary McDonald, W5UNF

RTTY test generator

The RTTY RY generator described in the March, 1971, issue of *ham radio* can *be made into a compact test unit,* including a 120-volt loop supply and 3.3 volt Vcc for the generator board, and housed in a 2-1/8 x 3 x 5¼-inch Minibox. The supply for the generator board is obtained by using a 6.3-volt transformer (Radio Shack 273-050) and bridge to supply approximately 9 volts dc. This is connected to an NPN emitter-follower voltage-regulator circuit. A 4-volt zener in the base of this transistor provides a regulated 3.3 volts at the emitter.

The 120-volt loop supply is added by connecting another 6.3-volt transformer back-to-back with the one used in the 3.3-Vdc supply. This provides 110 Vac, isolated from the power line, which is rectified to provide the 120-Vdc loop voltage. A 2500-ohm adjustable resistor permits setting loop current to 60 mA. Two closed-circuit phone jacks are connected in series with the loop supply. With the printer plug in one jack, loop current can be monitored with a milliammeter plugged into the other jack. Local copy can be generated by plugging the keyboard into the second jack and, of course, when the RY generator is active the loop is keyed to give local RY copy.

The clutch circuit is wired in this unit to a normally-closed momentary-contact switch. This provides a steady 60 mA of magnet current which may be keyed by the keyboard for local copy. Depressing the momentary contact switch will cause the loop to be keyed with a stream of RYs until the switch is released.

Tom Gibson, W3EAG

added uses for the vom/vtvm

Most of us are familiar with the normal uses of the vom or vtvm, and the scales on the instrument may occasionally give us other ideas. Perhaps you have

already used yours to check which end of an unknown diode was the cathode, or for the polariy of a transistor. Did you think of your vom as a current and voltage source?

Volt-ohmmeters (and vtvms, too) measure resistance as a matter of current flow through the meter with, usually, an internal battery as the source. A combination of meter shunts and series resistors give a number of ranges. Many of these, where related by x10, x100, etc., vary the current inversely by the same factors. A typical Rx1 range may have an external current flow of 50 mA, reducing to 5 and 0.5 mA on the Rx10 and Rx100 ranges. Since most 20,000-ohm-per-volt voms have a basic 50-μA movement, this is used for the Rx1000 range, sometimes shunted to a slightly higher value.

Some voms and vtvms use a single 1.5-V cell, others use two, three, four or more. This voltage is available between the test leads. Occasionally this voltage will be varied between the lower and higher ohm ranges. The internal impedance of the voltage/current source is indicated by the ohms reading at the center scale of the meter movement; i. e., the point where the meter movement is reading half of full-scale current or voltage, but indicated on the ohms range, multiplied by whatever range is in use. For example, on my Simpson 260, the center scale reading is 125 volts on a 250-volt scale. Immediately opposite this is a 12-ohm reading, which is the internal impedance on the Rx1 range. On the Rx100 range this will increase to 1200 ohms. Some variation will be found due to battery drain and condition.

I suggest that for ready use of the vom for a current or voltage source, that a table of values be applied with a piece of tape to some convenient spot on the instrument. This should include the polarity of the test leads for voltage between them, the voltage normally found on various ohm-meter ranges, and the current flow at full-scale reading (prods shorted). The internal impedance can be read directly off the ohmmeter scale.

The values found on seven more or less typical instruments in my shop are listed in table 1. They will give you an idea of what to expect. Here's hoping your vom or vtvm can be even more useful to you.

Eugene Hubbell, W7DI

table 1. Current flow of various volt-ohmmeters when used as a current source.

	Micronta 22-022	Simpson 260 series 5	Weston 980	Weston 564	Eico 104	Triplett 625-N	RCA Voltohmyst WV-87A 20
Rx1	50 mA	100 mA	60 mA	140 mA	60 mA	*70 mA	125 mA
Rx10	5 mA		6mA	14 mA	6 mA		14 mA
Rx100	.5 mA	1 mA	0.6 mA	1.4 mA		*15 mA	1.6 mA
Rx1K	50 μA		60 μA	.14 mA	60 μA		.16 mA
Rx10K		60 μA				*73 μA	15 μA
Rx100K							1.5 μA
Rx1000K							

*On 400, 50k and 10M ranges.

Micronta 22-022, 20k/v, 3-V battery
Simpson 260, 20k/v, 1½/6-V battery
Weston 980, 20k/v, 1½/15-V battery
Weston 564, 1K/V, 4½-V battery
Eico 104, 20k/V, 4½-V battery
Triplett 625—N, 10/20k/V, 4½-V battery
RCA Voltohmyst, 1½-V battery

counter reset generator

Rather than fuss around with discrete components while building the reset generator for a new counter, I decided to go modern and use the 74121. This 14-pin DIP IC really makes life easy as shown in fig. 3. My counter uses a 74196 in the units position and its clear demands a low to reset. The other four counters are 7490 decades which demand a high to reset, but must be kept low to count. The two outputs of the 74121 monostable fill

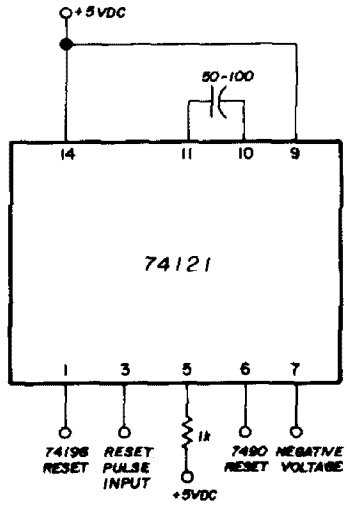


fig. 3. Counter reset generator circuit.

the bill quite nicely. The timing capacitor has to be large enough to produce an output pulse that will reset the 74196 reliably. If this is done, the 7490 decades will reset reliably also. The 74196 seems to be the fussier of the two counters, hence, this comment.

Allan S. Joffe, W3KBM

rigid mobile mount

I mounted a newly-purchased Japanese two-meter transceiver under the dash of the car with the one bracket supplied. The unit was not rigid enough and required another bracket from the rear of the unit to the heater duct. The extra bracket had to be designed so that the rig could be taken out easily. I solved the problem by means of the antenna connector on the rear of the transceiver.

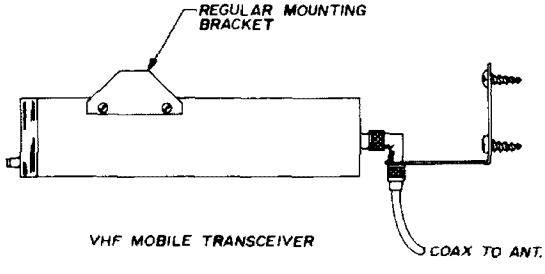


fig. 4. Bracket at rear of mobile transceiver greatly increases rigidity.

First, I screwed a right-angle connector (Amphenol UG-646) onto the chassis connector. I bent a right angle bracket, drilled a 5/8-inch hole in it and mounted the bracket so the right-angle connector fitted through the hole. The connector from the antenna was then screwed down, thus making a rigid mount.

I have found that the single bracket with which most of the small transceivers are supplied is not adequate for a rigid installation.

Vern Epp, VE7ABK

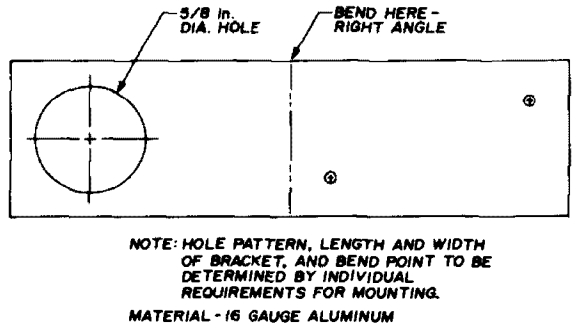


fig. 5. Layout for the rear mounting bracket.

restoring panel lettering

A white tire lettering stick (used by sports car owners to restore lettering on their racing tires) is ideal for restoring faded lettering and dial graduations on old equipment. The stick is a paint product and will be more durable when dry than other marking methods. Also, the sticks are easy to obtain at almost any tire or automotive store.

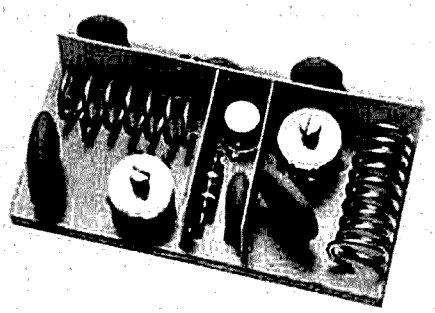
To restore an engraved plate, first be sure the panel and engraving are perfectly clean or the white will not adhere to the indentations. Cut a fresh surface on the end of the lettering stick and rub the stick over the engraving until the letters and lines are filled flush with the panel surface. Wrap a paper towel around something flat and immediately rub off the excess white paint. Do not wait to do this as the white will dry on the panel where it is not wanted.

On particularly bad surfaces, it will take two or more tries at cleaning and filling with the lettering stick. Be sure to start with a fresh surface on the end of the stick each time.

Ross N. Hayes, W8CL

new products

two-meter preamp

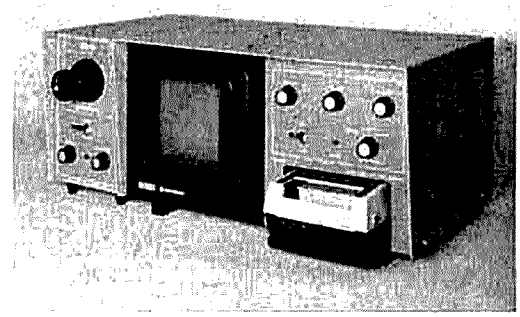


Boasting 20-dB gain and a 2.5-dB noise figure, Data Engineering's new two-meter preamplifier is designed to boost the sensitivity of most two-meter a-m, ssb and fm receivers to 0.1 μV. The tiny unit can be mounted inside of most existing receivers and transceivers or added externally in a minibox enclosure.

The unit uses a mosfet circuit with input, output, mosfet and power components separated and rf bypassed by gold-plated copper shields. With a 90-day guarantee, the unit sells for $9.50 in kit form and $12.50 fully assembled. Stock units are powered by 12 Vdc. An option for 150 to 250 Vdc operation is available for $2.95.

More information is available from Data Engineering, Inc., Box 1245, Springfield, Virginia 22151 or by using *check-off* on page 110.

slow scan tv system



Linear Systems, Inc., has announced the introduction of the SBE Scanvision system designed for use in the exciting and rapidly growing slow-scan television amateur radio market. Slow-scan television is undoubtedly the most exciting thing to happen in amateur radio in many years, and Linear Systems is very pleased to become involved in it at this relatively early stage. It brings the amateur hobby into an area of new technology which permits the transmission of pictures in signal bandwidth no greater than that required for voice signals, and represents a chance for amateurs around the world to get to know one another even better.

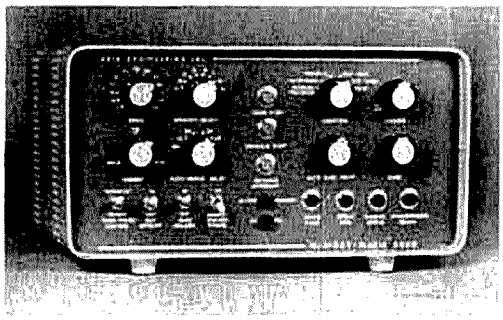
The two-component SBE slow-scan system consists of a specially designed high resolution tv camera designated model SB1-CTV. The other component in the system is the model SB1-MYV monitor. Controls on the monitor provide selection of camera, receiver or tape source of video signal, contrast, brightness and horizontal hold.

A completely unique feature of the Scanvision monitor not available on other equipment of this sort is a built-in cassette tape player which permits simultaneous recording of off-air signals for storage. It also permits the preparation of programming at more convenient times. The tape recorder can be removed from the set.

The list price for the camera, slow-scan monitor and integral tape recorder is $999.90. For more information, contact Mr. David C. Thompson, President, Linear Systems, Inc., 220 Airport Boulevard, Watsonville, California 95076, or use *check-off* on page 110.

More Details? CHECK—OFF Page 110

super-keyer



Perhaps the most sophisticated keyer on the market, Data Engineering's new Memory-Matic 8000 keyer features not just a very flexible iambic keyer, but the capabilities of eight 1000-bit, plug-in message memories. Messages in memory can be interrupted during transmission to insert impromptu additions — all without a discernible change between the transmitted code speed of the recorded message and the added comment.

Memories are plug-in with both 500- and 1000-bit memories available. Additionally, the unit can send any message in memory automatically at fixed intervals — as needed, for example, in meteor scatter communications. These intervals are clocked by the 60-Hz line current and can be synchronized to WWV.

A switch on the keyer can eliminate the dot, dash, or both memories with or without automatic character and work spacing — proving as fully automatic or as partially automatic operation as desired. The unit also has variable dot to dash weighting, two tune positions including "dot tune" for ssb transmitters, self-test and off-the-air testing capability, provision for use with many types of keys, a powerful built-in monitor and speaker and a built-in 117/220 Vac power supply.

The details on the flexibility and possibilities of this new keyer would take up too much space in this column. Full information is available by writing to Data Engineering, Inc., Box 1245, Springfield, Virginia 22151 or by using *check-off* on page 110.

The Memory-Matic 8000 sells for $398.50 with three 500-bit memories and one 1000-bit memory. Additional plug-in 500-bit memories are $21.50 and 1000-bit memories are $37.50.

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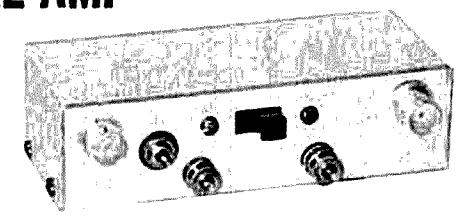
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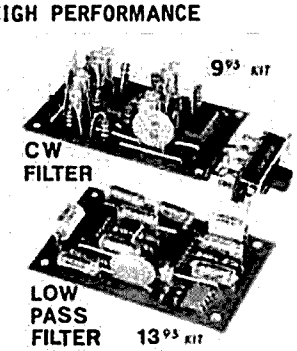
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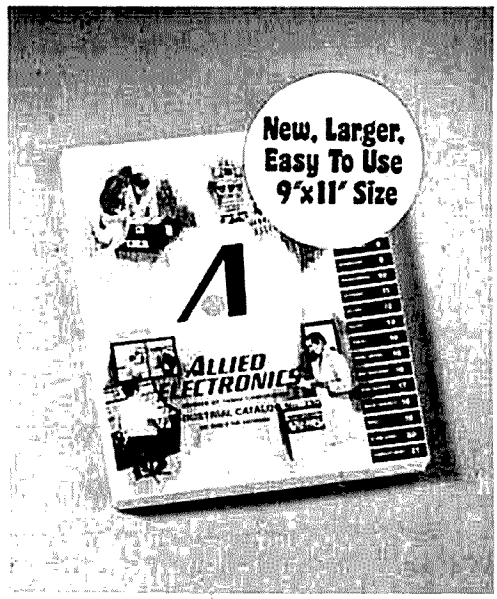
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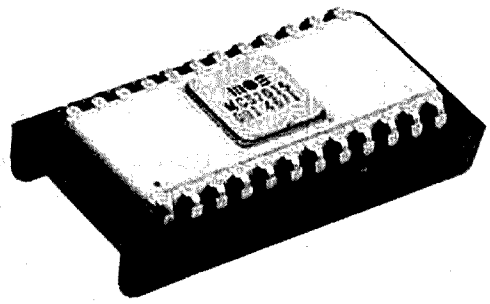
The new 1973 Allied Electronics catalog is a comprehensive buying guide for everything in industrial electronic parts and supplies. Compiled to meet the needs of industry, schools, institutions and government agencies, it's also the catalog for everyone looking for one dependable source for hard-to-get items — one or a thousand. Considered by many as the "Bible of the Industry," the catalog lists over 50,000 separate stock items from more than 400 manufacturers. Merchandise is grouped by sections; numerical cover margin tabs guide you quickly to the products you need.

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Other major sections include test equipment, intercoms, power supplies, electronic counters, sound equipment, chemicals, hardware, technical books, tools and solder equipment. Allied Electronics Catalog no. 730 sells for $5.00 (or *free* with $10.00 minimum order), and is available from Allied Electronics, 2400 W. Washington Boulevard, Chicago, Illinois 60612.

More Details? CHECK—OFF Page 110

time zone converter



Southwest Technical Products Corporation has developed a read-only-memory time zone converter using a TTL compatible MOS circuit in a standard 24-pin ceramic package. This ROM converts the input time zone information used in digital electronic clocks into any other selected twelve-hour time zone. The addition of the time zone converter makes possible instant switch selection of any world time zone on any digital clock using standard BCD logic for the readout system. The 4671 automatically takes care of borrows, carries and other necessary logic to make the desired conversion. Normal TTL logic levels, +5 V and -12 V are required for operation. Single unit price is $19.50. More information is available from Southwest Technical Products Corporation, 219 West Rhapsody, San Antonio, Texas 78216 or from *check-off* on page 110.

ham radio film

Media Five, a West-coast film distributor, has just released a new 15-minute, 16-mm color motion picture entitled, "This is Ham Radio". As well as demonstrating practical science in action, "This is Ham Radio" details the fun, challenge – and real importance – of a fascinating hobby shared by more than a half million radio amateurs around the world.

The film stresses licensing requirements, low-cost rigs, the relative ease of earning a Novice ticket from the FCC and getting on the air. Fast-paced scenes show dozens of different activities, and featured at every turn are young people: operating mobile rigs, testing emergency

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Untested TO-5 Transistors. All with full leads. Some NPN some PNP some Silicon some Germanium. Spot check shows about 60% good units. 100 for $3.75 ppd.

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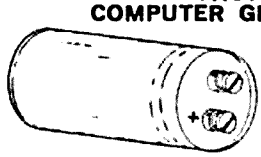
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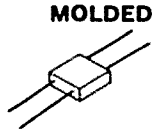
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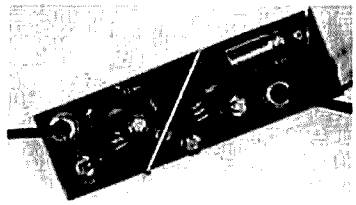
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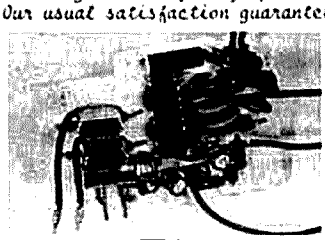
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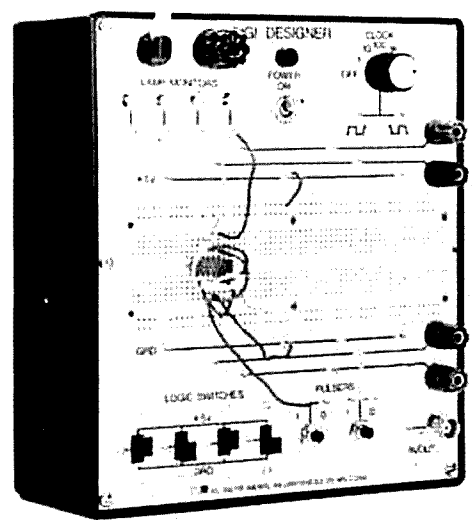
GREENVILLE, NH 03048

gear (in the rain, of course), relaying messages across the ocean or just rapping with friends across town.

Narrator is college-bound teenager Matt Futterman, WB6KPN, Extra-class ham who got his first license at age 13. He describes some of the lure of ham radio: building your own equipment, assisting Civil Defense officials in emergency communications, meeting new friends, learning the Morse code well enough to think in it, sharing the excitement of world-wide person-to-person communication and getting involved in something really worthwhile.

"This is Ham Radio," educational counterpart of the awarding winning "The Ham's Wide World," was produced in cooperation with the American Radio Relay League. "This is Ham Radio" is available for rental, purchase, or preview toward purchase from Media Five Film Distributors, 1011 North Cole Avenue, Hollywood, California, 90038. Sale $175. Rental information on request. Produced 1970.

digi-designer



EL Instruments announces a new aid in breadboarding for the experimenter interested in digital circuits. The Digi Designer incorporates the EL SK-10 component socket, a variable, six position 1 to 100 kHz clock, four logic lamps, four switches, two bounce-free pushbuttons

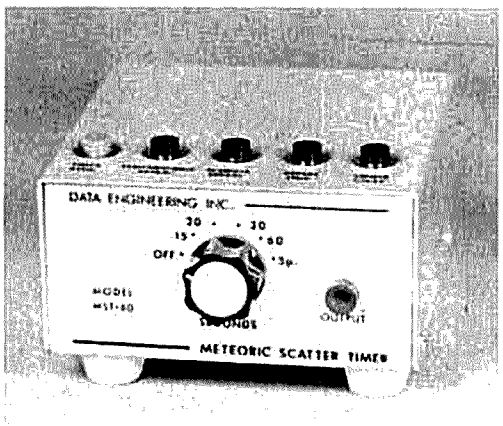
More Details? CHECK–OFF Page 110

for use as pulsers and an internal 5 Vdc power supply. There are numerous terminal points on the front panel for external inputs or patch cords.

The Digi Designer enables the user to completely design and test a circuit by merely plugging his components into the SK-10 socket, and interconnecting with standard number 24 AWG hook-up wire. No soldering is necessary.

The Digi Designer is available from stock in kit form for $49.95 or wired for $95.00. Complete information is available from EL Instruments Inc., 61 First Street, Derby, Connecticut 06418 or by using *check-off* on page 110.

communications timer



Data Engineering has introduced a precise timer for use with meteor scatter, moon bounce or tropospheric scatter communications. Transmissions can be synchronized with WWV and transmitted at precise intervals of 15, 20, 30 and 60 seconds. The output of the timer is used to automatically start coded CW transmissions, while the output indicator signals the start of manually sent CW.

The unit designated the MST-60, sells for $49.50 and carries Data Engineer-protection against rf and line spikes affecting the timer's accuracy.

The unit designated the MST-60, sells for $49.50 and carries Digital Engineering's standard five year guarantee against defects in parts and manufacture. More details are available by writing to Data Engineering, Inc., Box 1245, Springfield, Virginia 22151 or by using *check-off* on page 110.

Reception

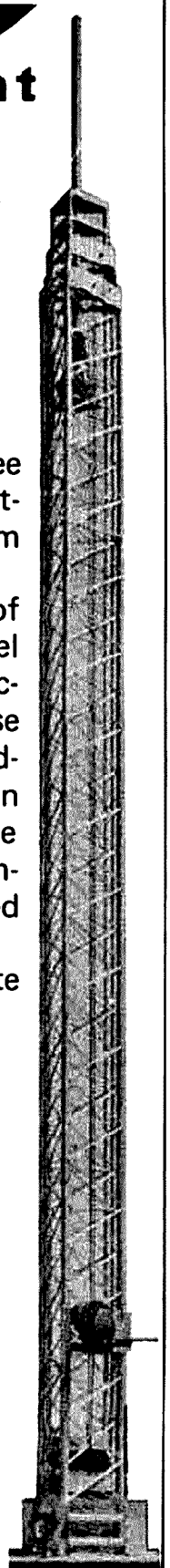
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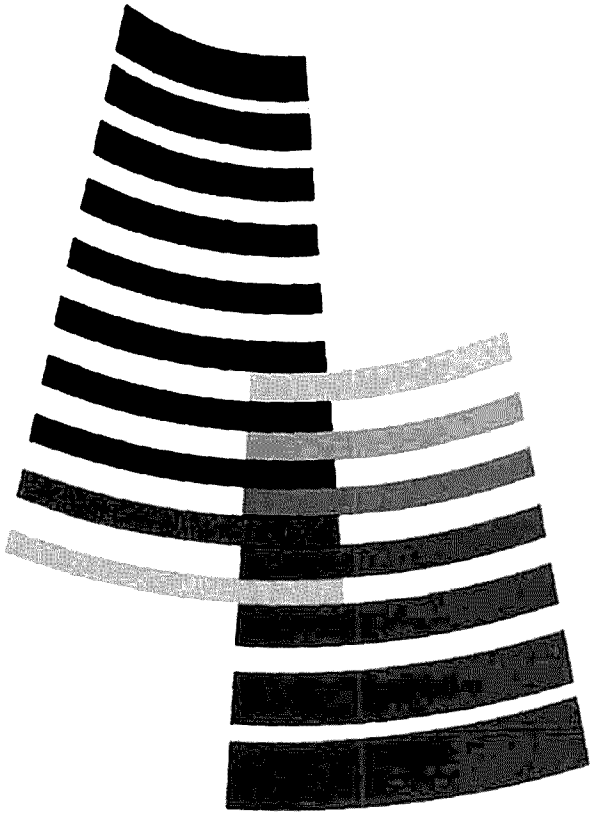
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- vhf fm receiver scanner 22
- solid-state noise blanker 38
- tunable audio oscillator 50

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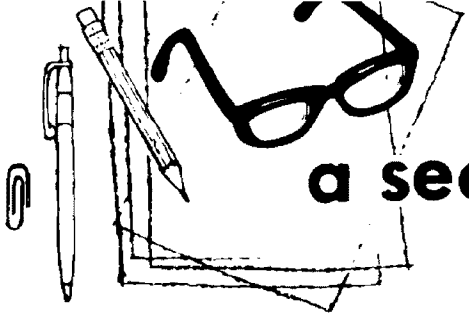
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a second look

by jim fisk

A little over twenty-five years ago, on December 23rd, 1947, to be exact, a group of scientists at Bell Laboratories built a one-stage amplifier circuit around the world's first transistor, giving birth to a whole new era of electronics and communications. But the beginning of the story was not in 1947, but long before. There had been hints of amplification in semiconductors as early as the 1920s but few experimenters could duplicate the results. Nobody realized the effect of semiconductor impurities nor understood the action of semiconductor materials.

In 1930 Dr. Julius Lilienfeld, a German physicist, actually patented a semiconductor amplifier that could be compared to today's mosfet. Although Dr. Lilienfeld's amplifier worked, it could not be duplicated by other workers, and it slowly slipped into oblivion.

In 1939, Dr. William Shockley made an entry into his lab notebook at Bell Labs, "It has today occurred to me that an amplifier using semiconductors rather than vacuum is in principle possible." It was nearly eight years before this concept would bear fruit. A large part of this period was spent in learning more about that old bugaboo, semiconductor impurities.

The 1N21 crystal detector, developed during World War II and the workhorse of wartime radar receivers, provided some of the impetus. After the war a solid-state research team at Bell Labs, co-headed by Dr. Shockley, started experimenting with germanium and silicon, two semiconductors that were easy to work with. As one of the group says today, "We felt that the area was so fertile that you could devise an experiment in the morning, go out in the lab and try it in the afternoon, and then write a paper about it that evening."

The first device the group attempted to build was what is now called an insulated-gate fet. The device didn't work. The group scrambled around, dug into the literature and spent long hours discussing the alternatives.

Dr. Walter Brattain tried an experiment where he covered a metal point with a thin layer of wax and pushed it down on the surface of a piece of silicon. He then surrounded the point with a drop of water and made contact to it. The water was insulated from the point by the wax layer. He found that voltages applied between the water and the silicon would change the current flowing from the silicon to the point. Power amplification had been achieved! Unfortunately, the drop of water would evaporate almost as soon as things were working well.

This led to experiments with other electrolytes that didn't evaporate so readily. Then, they discovered a thin oxide layer on the surface of the semiconductor under the electrolyte and decided to use a spot of gold as a field electrode to eliminate the electrolyte.

When this was tried, an electrical discharge between the point and the gold spoiled a spot in the middle – when they had washed off the electrolyte they had inadvertently washed off the oxide film, which was soluble in water. However, by placing the point around the edge of the gold spot they observed a new effect – when a small positive voltage was applied to the gold, holes flowed in the surface of the semiconductor, greatly increasing the flow of current. Four days later two gold contacts less than two-thousandths of an inch apart were made to the same piece of germanium and the first transistor was born.

Nine years later, in 1956, the three inventors, Dr. William Shockley, Dr. John Bardeen and Dr. Walter Brattain were awarded the Nobel prize in physics. Little did they realize that their crude laboratory device would spawn a multi-billion dollar semiconductor industry that today affects all our lives.

Jim Fisk, W1DTY
editor

designing communications receivers for good strong-signal performance

Many modern amateur receivers perform poorly in the presence of strong signals — here are some sound design ideas for solving this problem

Ray Moore, 116 Walnut Street, Walpole, Massachusetts 02081

"Current receivers . . . perform poorly under exactly those conditions that are most important, when the desired signal is weak and the undesired signal is strong" is the way Squires began an article on receiver front-ends in 1963.[1] Since then, solid-state receivers have become the standard in new design, increasing the difficulty of obtaining adequate signal handling characteristics. Outstanding strong-signal performance is one of the more difficult and expensive characteristics to design into a receiver; the least expensive characteristics are sensitivity and gain. This is why most receivers are "hot" enough to rattle the walls on noise alone.

The price for low noise figures on the high-frequency bands is paid, not in dollars, but in poor to mediocre strong-signal performance. Top professional receivers in the $5000 to $10000 price range are often designed for noise figures of 10 to 12 dB, the excess noise-figure performance being traded for signal handling ability.

What are the symptons of poor strong-signal performance? Under actual on-the-air conditions a receiver which cannot handle strong signals can have its performance so deteriorated that its static sensitivity and selectivity figures are meaningless. The receiver may go completely silent in the presence of a strong, unwanted station, or its gain and sensitivity may be so reduced that you can not copy the weak, desired station.

The amateur bands may appear to

be full of weak commercial stations which are really operating *outside the* bands — heterodynes, birdies and broadcast stations appear. Strong ssb stations produce less obvious effects which make the bands seem noisy and full of splatter. All these conditions are produced by deficiencies *within* the receiver.

There are a number of ways in which strong, undesired signals outside the i-f passband can interfere with reception. Some require only a single undesired signal to be present; others require two or more signals. Single signal effects include adjacent channel interference, image interference, i-f breakthrough, strong-signal spurious responses, desensitization and blocking. Multiple signal effects include cross-modulation and rf intermodulation.

single-signal interference

Adjacent-channel interference is caused by a strong, undesired signal close to, but outside of, the receiver's i-f passband. This is the fault of the main selectivity determining filter. The filter either has a poor shape factor or the ultimate attenuation (stop band) is not deep enough. Some filters, for instance, have skirts that go down to only 50 or 60 dB before they flatten out. One solution to this problem is cascading two or more filters. Care in matching filter center frequencies and proper isolation produce ultimate attenuations of 120 dB or more and shape factors approaching 1:1.

Image frequency interference is a function of rf selectivity and the frequency of the first i-f. Secondary images between the first and second i-f stages of multiple-conversion receivers are also possible. Many current receivers have image rejection ratios as low as 50 dB on at least one band. However, with an i-f in the megahertz range, image ratios of 100 dB or more are possible if economy is not an overriding factor. Up-conversion to an i-f above the receiver tuning range in conjunction with a front-end low-pass filter is also very effective.

I-f breakthrough occurs when a signal on the frequency of the i-f rides through the front-end and into the i-f by brute force. The i-f rejection ratio of a receiver is a function of its rf selectivity, and runs in the range of 50 to 70 dB for amateur receivers and up to 70 to 100 dB for professional and military receivers.

The problem is much more severe with *variable* first i-f receivers than with *fixed* first i-f receivers because it is difficult to find a 500- to 1000-kHz slice of the spectrum with no strong stations, and further, because simple traps can reduce i-f breakthrough in fixed i-f receivers where band elimination filters are required for variable i-f receivers. Up-conversion with a low-pass filter is one of the most effective solutions to the i-f breakthrough problem.

Strong-signal spurious responses are produced in two ways. First, strong signals can ride through the front end and mix with harmonics of the first oscillator to produce the i-f. Second, a nonlinear rf amplifier or mixer can generate harmonics of a strong signal which beat with the local oscillator. Rf selectivity and filtering of the injection frequencies are the cure for this problem.

The best receivers have strong-signal spurious responses of 100 dB in relation to a 1-μV desired signal. That is, it takes a 100-mV undesired strong signal to produce a receiver output equivalent to the output produced by a 1-μV desired signal. A 0.3 mV signal (50 dB relative to 1 μV) can produce strong signal spurious responses in less expensive receivers.

Desensitization and blocking are different degrees of severity of the same problem. In desensitization a strong signal outside the passband reduces the gain of the receiver which may make it impossible to hear a weak desired signal. In blocking, the gain is reduced to such an extent that the receiver goes silent. Both desensitization and blocking are caused by a signal which rides through and is rectified by the first active device causing a shift in the operating point of the device. If the first device is also connected

to the agc system, the rectified voltage can be fed back through the agc line to affect other stages as well. These problems are most common in the immediate vicinity of transmitters.

Professional class receivers specify desensitization and blocking in terms of the unwanted signal level required to cause a 3-dB drop in receiver output when tuned to a 1 mV desired signal. 100 mV is a typical desensitization specification for such a receiver. These characteristics are seldom specified on lower priced receivers but tests have shown that some will completely block at 35 mV or less.

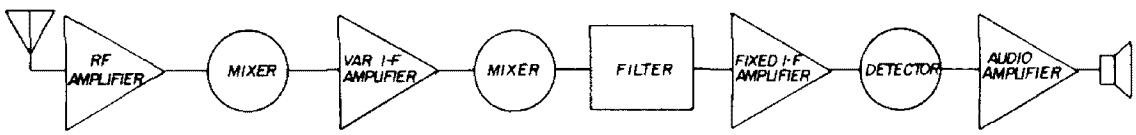


fig. 1. Signal path through a typical modern communications receiver.

cross-modulation

Cross-modulation is a mixing effect that is produced when a desired signal and a strong, undesired signal are applied simultaneously to a device with third-order curvature of its input-vs-output transfer characteristic. The first and second mixers are the stages most likely to produce cross-modulation but very strong signals can also produce it in the rf amplifier. The result of cross-mod is that the modulation of the undesired signal is superimposed on the desired signal and cannot be removed by subsequent processing. Cross-mod is aggravated in the first stage of a receiver by the popular gain control methods which change the operating conditions of the device (see **fig. 5**). Any attenuation ahead of the offending stage is beneficial even if the desired signal is also attenuated, because 1 dB of attenuation reduces cross-modulation by 2 dB. Of course, attenuation of the undesired signal by rf selectivity without affecting the desired signal is preferred, but this is not always possible if the undesired signal is close to the passband.

Cross-modulation is seldom specified except in professional receivers and even there a lack of standard conditions and levels makes direct comparisons impossible. The undesired signal required to produce a certain level of cross-mod is specified, but the reference cross-mod level is variously given as -10 dB, -20 dB or -30 dB relative to the desired signal output, as 3% cross-mod, or as "negligible". More important, the spacing of the undesired signal from the desired signal may be anywhere from 10 to 100 kHz or may be expressed as a percentage up to 10%. Cross-mod performance is improved by use of greater rf selectivity and more linear active devices. A given active device can often be made more linear by optimizing its operating conditions.

intermodulation

Rf intermodulation (IM) is, like cross-modulation, the result of third-order curvature of a device. If there are two strong undesired signals, f_1 and f_2, they will produce two third-order products, one at $f_1+(f_1-f_2)$ and the other at $f_2-(f_1-f_2)$. Fifth-order curvature can also produce much weaker IM products spaced $2(f_1-f_2)$ from the two offending signals. As an example, two strong signals, one at 3450 kHz and the other at 3490 kHz, can produce two IM products in the 80-meter band, a third-order product at 3530 kHz and a fifth order product at 3570 kHz. Two other IM products fall outside the amateur band at 3410 and 3370 kHz. Perhaps this is why the 80 meter band on

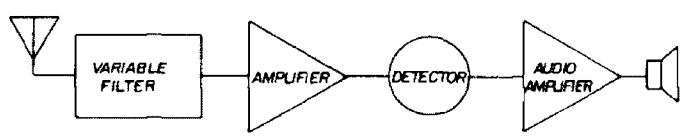


fig. 2. Ideal receiver signal path for handling strong signals.

your receiver sounds full of RTTY and other commercial stations.

There are two ways in which IM is specified in professional receivers. One is to tell how far the IM product is below the level of the two signals which cause it. Third-order IM, for example, can be specified as, "at least 60 dB below two 10-mV undesired signals." IM can also be specified as the undesired signal level to produce an IM product equivalent to a 1-μV desired signal. The better professional and military receivers require from +70 dB to +100 dB relative to 1 μV to produce third-order IM equivalent to a 1-μV desired signal. Improvement in IM performance is accomplished by the same means as for cross-modulation.

receiver signal path

The typical modern communications receiver (**fig. 1**) has a signal path which is not ideal from the strong-signal standpoint. The problem with this design is that four stages are exposed to strong undesired signals before the passband-determining filter can reduce them to harmless proportions.

The ideal receiver configuration for handling strong signals (**fig. 2**) is a throwback to some of the TRF receivers of the 1920s which had all their tuned circuits lumped between the antenna and the first rf stage. The filter in the ideal receiver, however, must be adequate to provide adjacent channel selectivity to modern standards – a shape factor of 2:1, 2.5 kHz bandwidth, ultimate rejection greater than 100 dB, insertion loss 1 or 2 dB. Such a filter would stop the strong undesired signals before they got into the receiver, unlike the two or three tuned circuits in most receivers which allow signals 50 or 100 kHz off frequency to ride through with little attenuation. High-frequency crystal-lattice filters which meet these requirements for single frequencies are available, and receivers such as this, covering one or several fixed frequencies, are in use.

A practical, tunable receiver which can approach the ideal in performance is shown in **fig. 3**. The objective is to provide maximum adjacent-channel selectivity as close as possible to the antenna. The mixer must be low noise and as linear as possible; FL1 should consist of two to four tuned circuits, depending upon the noise figure of the mixer and the target noise figure for the entire receiver. The i-f must be high enough so FL1 can adequately suppress images and low enough

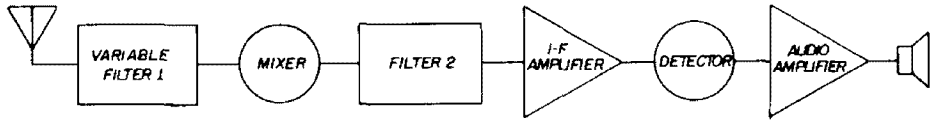


fig. 3. Compromise receiver signal path for handling strong signals.

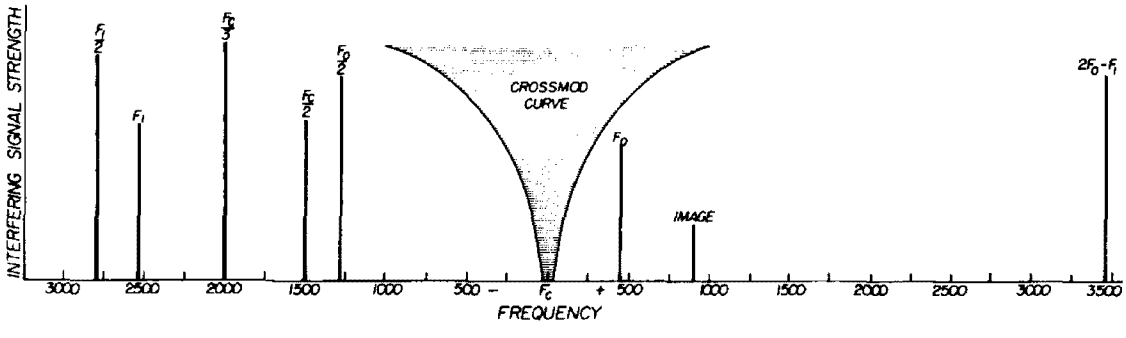


fig. 4. Plot of strong-signal responses in a typical communications receiver. Frequency f_c is the center frequency to which the receiver is tuned (3000 kHz), f_o is the oscillator frequency (3455 kHz), and f_i is the intermediate frequency (455 kHz).

to be practical, say, in the range of 1.5 to 50 MHz.

Higher intermediate frequencies remove the i-f from the receiver tuning range and permit the use of up-conversion in conjunction with a low-pass filter ahead of FL1. The disadvantage of the higher frequencies is that filters, while available, may be expensive and difficult

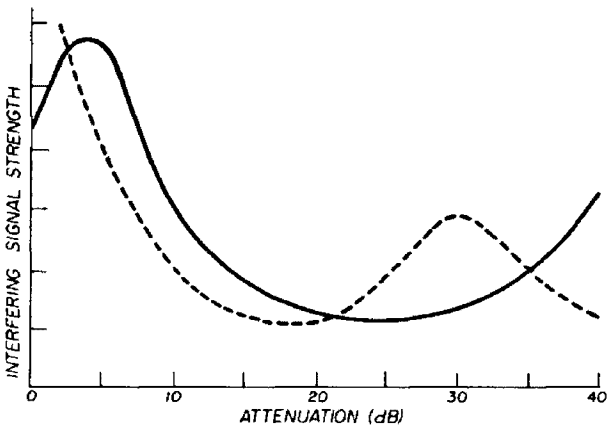


fig. 5. interfering signal level required to produce 1% cross-modulation in two different active devices.

to locate. The stability and constant tuning rate usually associated with the variable first i-f, multiple-conversion receiver can be achieved with this configuration by deriving the injection frequency for the first mixer through the pre-mixer technique. In this case, complexity has been removed from the signal path and added to the frequency determining circuits where it can do no harm to the signal.

receiver design trends

The receiver design trends of the last twenty years have, almost without exception, been in the direction of poorer strong-signal performance. One of the innovations of that period is all solid-state receiver design. In general, semi-conductors do not handle strong signals as well as vacuum tubes, and the use of semi-conductors in stages ahead of the adjacent-channel selectivity filters degrades strong-signal performance.

Multiple conversion, another design innovation, results in two mixers and three to five total stages before there is appreciable selectivity to protect the amplifiers and mixers from strong undesired signals. Broadband and variable i-f stages are susceptible to i-f breakthrough.

Furthermore, the newer low-noise devices, both tubes and semi-conductors, tend to be less linear than the older vacuum tubes. The low-impedance (50-ohm), tightly-coupled primaries on modern antenna coils compound the strong signal problem because they degrade the Q of the tuned circuit. They also present higher signal levels to the active device than did the old medium-impedance (200- to 500-ohm) primaries.

evaluating strong-signal performance

The best method of evaluating the strong signal performance of the receiver is that used for military and professional receivers. Two signal generators are used to plot cross-mod and spurious responses as shown in fig. 4. In plotting the cross-mod curve one generator feeds in a desired signal (10 to 100 μV) at the frequency to which the receiver is tuned; the second generator is swept out from the desired signal frequency, first in one direction and then the other, representing a strong undesired signal. The amplitude of the undesired signal necessary to produce the reference cross-mod level is recorded at enough points to produce the curve shown in fig. 4.

The first generator is removed from the circuit and the second generator is swept through the spectrum again. Any

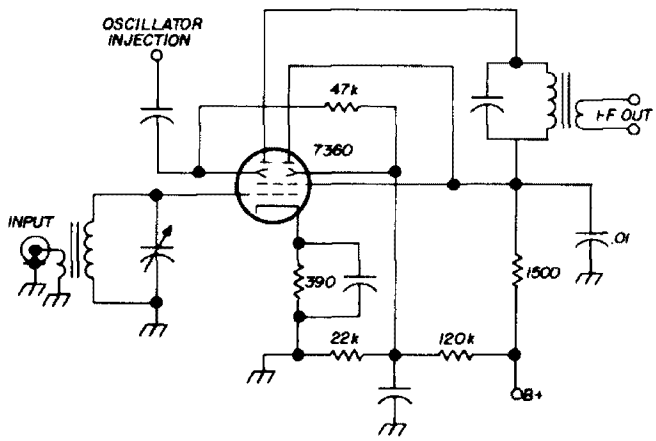


fig. 6. Unbalanced 7360 beam-deflection mixer circuit.

discrete responses such as image and i-f breakthrough are recorded. The undesired signal strength required to produce a response equivalent to a 1-μV desired signal is recorded for the discrete responses.

Since there are an infinite number of signal combinations which can produce IM, you must be content with spot checks in evaluating this characteristic. Set one generator at f_c plus 30 kHz and the other at f_c plus 60 kHz. The amplitude of both generators must be the same. Increase the amplitude of the two generators until a response appears at f_c equivalent to that produced by a 1-μV desired signal. The level of either generator is then the IM response level and may be expressed in dB relative to 1 μV.

There are only two ways to improve the ability of a receiver to handle strong signals: Reduce the strength of the undesired signals or, in the case of cross-mod and IM, improve the linearity of mixer and amplifier devices in the stages ahead of the adjacent-channel selectivity filter.

Reducing the strength of the undesired signals is the method used in the receiver configuration in **fig. 2.** If the undesired signals can't get into the receiver, they can't do any damage. In most receivers you have to get along with something less than ideal rf selectivity, but strive for the maximum practical selectivity. Reducing the undesired signal by 1 dB will reduce cross-mod by 2 dB and IM by 3 dB, so even a small amount of additional selectivity can make a decided improvement.

If ideal rf selectivity isn't available then the linearity of the amplifier and

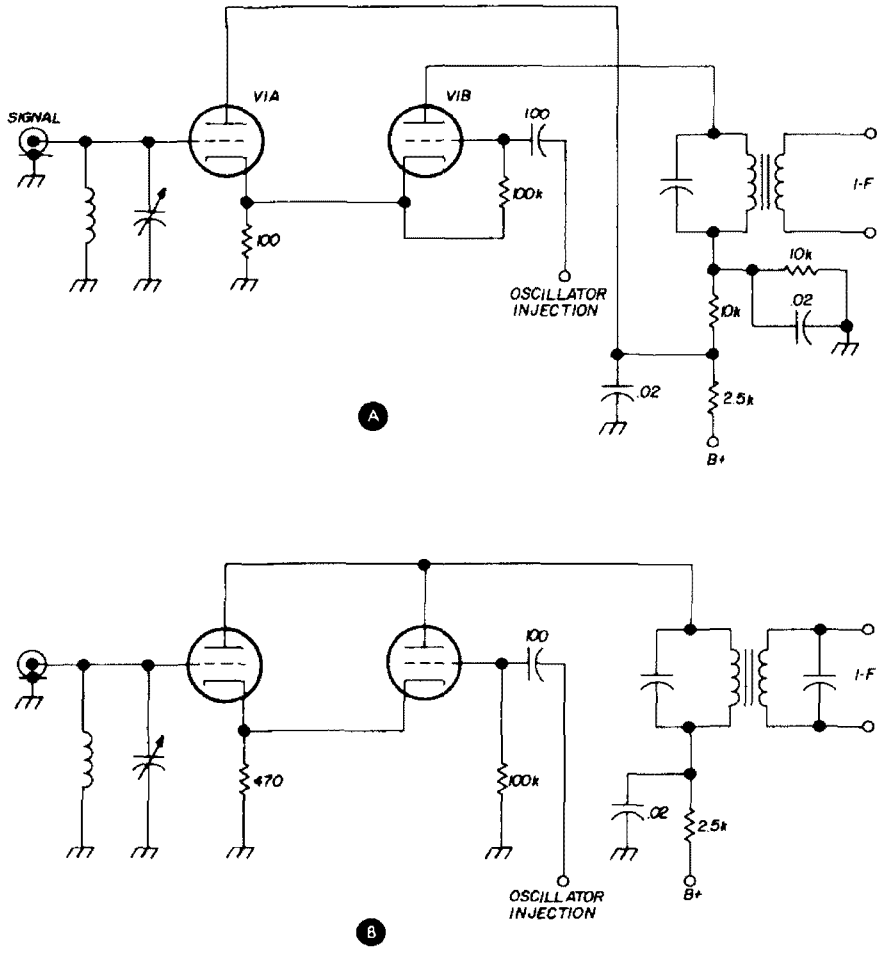


fig. 7. Two dual-triode mixer circuits. The circuit in (B) is not as sensitive as the circuit in (A), but (B) will handle larger signals. These circuits are suitable for vacuum tubes such as the 12AU7, 12AT7 or 6DJ8.

mixer devices becomes important. Not only should the most linear devices and circuits be selected, they must be optimized for strong-signal performance. Each active device has an optimum point at which it is least susceptible to cross-mod and IM. This is illustrated in **fig. 5**

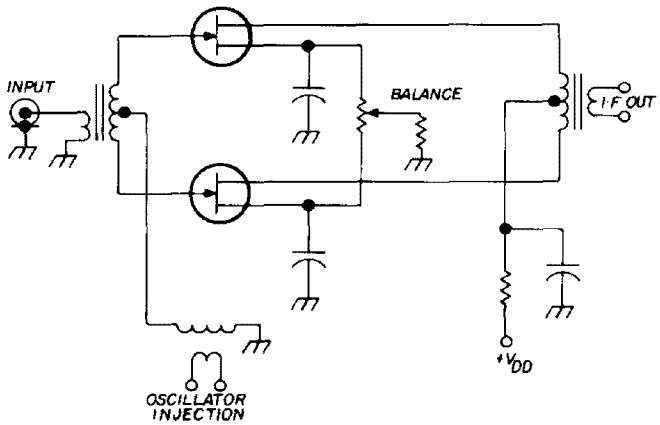


fig. 8. Balanced mixer circuit using a field-effect transistor.

which shows generalized cross-mod-vs-attenuation curves typical of active devices when attenuation is produced by changes in bias.

Most amateur receivers suppress all strong-signal effects from out-of-passband signals by a minimum of 50 dB. The suppression of military and professional receivers runs from 70 to 100 dB or more. The noise figure of amateur receivers runs from 5 to 8 dB, that of professional receivers is usually around 10 dB. The cost of amateur receivers is $250 to $800, professional receivers $5000 to $10,000 and up. Obviously, sensitivity is cheap while strong-signal performance is expensive.

It can not be stated too often that good strong-signal performance in modern receivers is the result of painstaking design work. It is not enough to pick out a signal path configuration and the circuit for each stage. Each stage must be experimentally optimized for operating point, voltages and injection levels. If there are two or more circuits under consideration for a given stage, each must be optimized, and then compared.

strong-signal performance

The following paragraphs detail the steps which are necessary in the design of a receiver with superior strong-signal performance.

First of all, determine the maximum acceptable noise figure. Many professional receivers are designed for a noise figure of 10 to 12 dB. If the input device of the receiver has a noise figure of 6 dB the designer then has an excess of 4 to 6 dB to use in providing additional rf selectivity, to keep the gain low ahead of the adjacent channel selectivity filter, and to allow optimum strong-signal biasing of stages.

For amateur work where it is possible to take advantage of the rare occasion when a 5- or 6-dB noise figure is usable the best idea is to have an auxiliary low-noise preselector that can be switched in ahead of the receiver. It is easy to add a preselector to a receiver that has sacrificed noise figure for superb strong-signal characteristics, but it is difficult to improve the strong-signal characteristics of a receiver that has been designed for low noise.

Next, select the signal-path configuration. For general use the arrangement of **fig. 3** is recommended. A low-gain rf amplifier can be added if the mixer does not have the required sensitivity. Note that it is *not* necessary to go to the variable first i-f approach to achieve the vfo type tuning we expect in modern receivers.

mixers

The mixer stage is the key to the design of a superior strong-signal receiver. If the configuration of **fig. 3** is to be used to its fullest, the mixer must provide low noise, high conversion transconductance and exceptional linearity. Fortunately, there are devices and circuits which possess these characteristics to a much greater degree than the mixers found in the average medium-priced receiver.

The problem with many mixers is that the device is biased to an operating point where it is not linear so that the required

second-order (sum and difference) frequencies are generated. However, at that point the device is also likely to be an efficient generator of third- and higher-order products which cause cross-modulation and rf intermodulation. Following are some recommended mixers which have been used in professional or military receivers or which are of unusual interest.

Squires developed a mixer circuit with the 7360 beam deflection tube in 1963 which has been used in a number of amateur receivers and at least one commercial receiver.[1] An unbalanced version of the circuit is shown in **fig. 6.** The tube is set up for linear operation between G1 and the plates; the signal is switched between the active plate and the grounded plate by the oscillator voltage appled to deflection plate.

Another source indicates that superior performance is obtained from the circuit with a fixed bias (about 1.9 V) and with an oscillator injection of 7.5 volts.[2] Squires' original balanced circuit had a noise figure of 5.5 dB at 29 MHz and handled undesired signals up to 1.5 volts.

Various dual-triode mixer circuits perform well because they are low noise and require little or no gain ahead of them. The 12AU7, 12AT7 and 6DJ8 are some of the tubes which have been used.

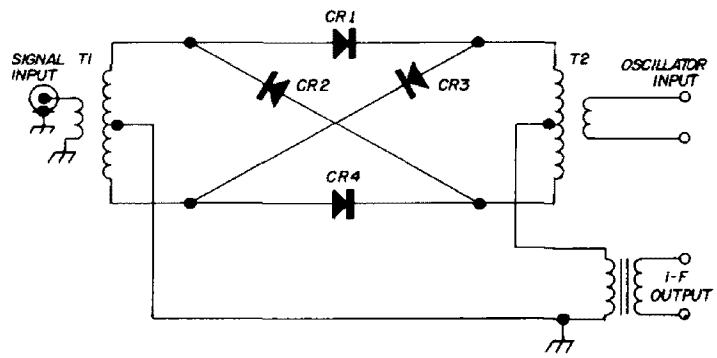


fig. 9. Diode balanced mixer has approximately 8-dB noise figure and handles large signals well. Diodes CR1-CR4 are hot-carrier diodes; transformers T1 and T2 are toroidal types.

In the circuit of **fig. 7A** V1B is a triode mixer with the signal injected at the cathode and the oscillator at the grid. V1A is a cathode follower which provides isolation between the signal and oscillator circuits. The 10k resistors in the plate circuit of V1B form a voltage divider to reduce the plate voltage because V1B must have a much lower transconductance than V1A. A somewhat different circuit is shown in **fig. 7B.** Sensitivity of this circuit is not as good as that of **fig. 7A** but it will handle larger signals.

The fet balanced mixer (**fig. 8**) is a good choice for use in the circuit of **fig. 3.** When used with suitable devices (such

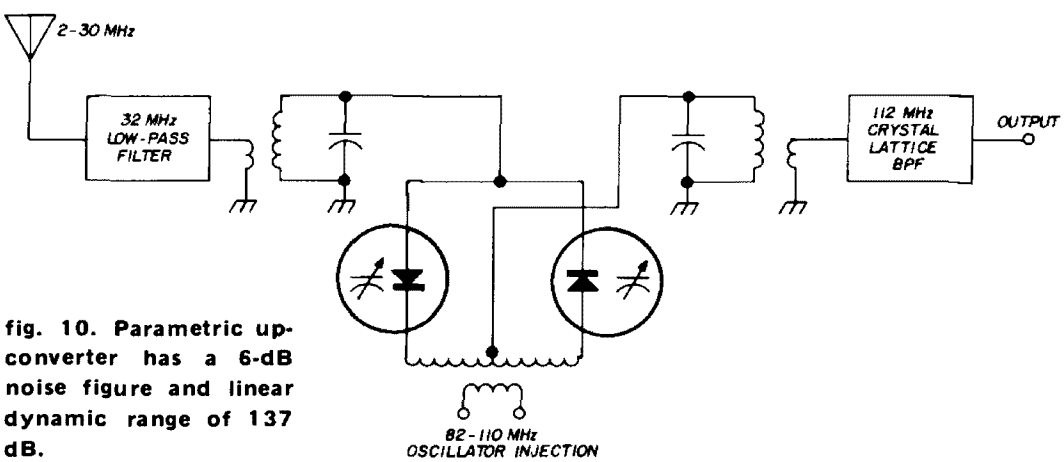


fig. 10. Parametric upconverter has a 6-dB noise figure and linear dynamic range of 137 dB.

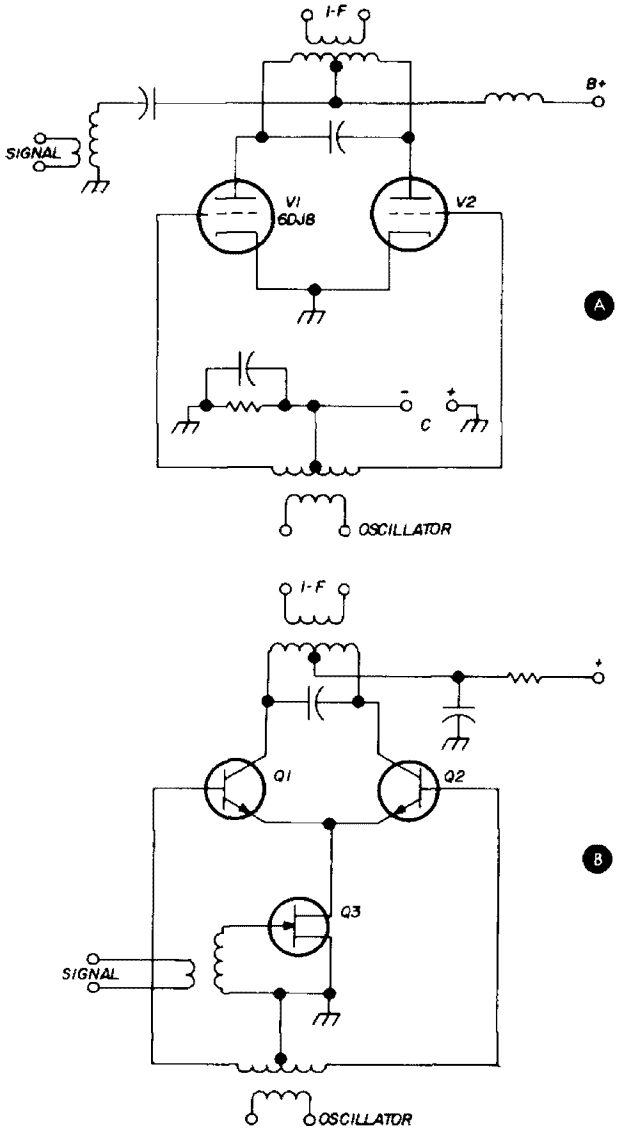


fig. 11. Vacuum-tube (A) and transistor-type (B) switching mixer circuits.

as the 2N4416) it has a noise figure of 4 to 5 dB and handles large signals fairly well. The noise figure is low enough to allow a rather elaborate rf filter to be placed between it and the antenna.

The diode balanced mixer using hot-carrier diodes in fig. 9 has a noise figure around 8 dB and handles strong signals well. However, it requires high local-oscillator power level and has conversion loss, so it must be preceded by an rf amplifier in order to achieve an acceptable overall noise figure. These are both disadvantages.

A parametric up-converter (fig. 10) is used in the National R1490/GRR17 wide dynamic range military receiver.[3,4] The specified linearity of the circuit is 137 dB and the noise figure is 6 dB. All external spurious responses, images, i-f breakthrough, cross-mod, IM, etc., are down more than 100 dB. The 112-MHz filter sharply attenuates all undesired signals more than a few kHz outside the i-f passband. Adjacent-channel selectivity is provided in a later stage at 5 MHz.

Up-conversion is advantageous because a simple low-pass filter can be used to increase the rejection of the i-f and image frequencies without the added complexity or loss that additional tuned circuits would bring. The up-conversion technique can be used with any of the other mixer circuits as well.

The 7360 mixer operates as a very linear switching circuit. Attempts have been made to improve the technique and to develop semiconductor equivalents. Perhaps the ultimate performance so far was also by Squires with a dual-triode switching mixer (fig. 11A) in a receiver that handled 3-volt rms signals within 10 kHz of the desired signal, and 25-volt signals 10% removed, without cross-mod.

He used a 6DJ8 tube and the oscillator injection had a square-wave characteristic. The tube was biased as a class-C amplifier and the oscillator had sufficient amplitude to drive the grids positive. The signal was alternately switched to ground through the two tubes by the oscillator injection voltage. A stage of rf amplification was used ahead of the mixer which worked into a 6-kHz wide filter. A somewhat similar scheme using semiconductors (fig. 11B) has been tried, but details on its performance are not available.[5] The signal is amplified in Q3 and alternately switched to ground through Q1 and Q2 and the output transformer.

rf amplifiers

If an rf amplifier must be used, the rule is to use only enough gain to override the mixer noise, using any excess gain to provide additional rf selectivity either by adding tuned circuits or tapping down on the existing tuned circuits to improve Q. In general, the best rf amplifier devices are vacuum-tube pentodes or triodes in the cascode circuit. Next in order are fets,

and last are bipolar transistors. When receivers *must* be all solid-state, circuit designers have gone to power fets and power bipolars for rf amplifiers in order to improve strong signal performance.[3,6]

Rf amplifiers must be adjusted experimentally to determine the best bias, plate, screen, drain or collector voltages. If the gain is higher than required it should not be reduced by changing the operating conditions, but by adding rf selectivity through additional tuned circuits, looser coupling or tapping down.

gain control

The method by which the front-end gain is controlled is important to strong-signal performance. As shown in **fig. 5**, if a conventional gain control method is used which changes the operating point of the device, it results in serious degradation in the ability of the device to handle strong signals. The preferred method of manual gain control is a resistive attenuator ahead of the first active device (**fig. 12**).

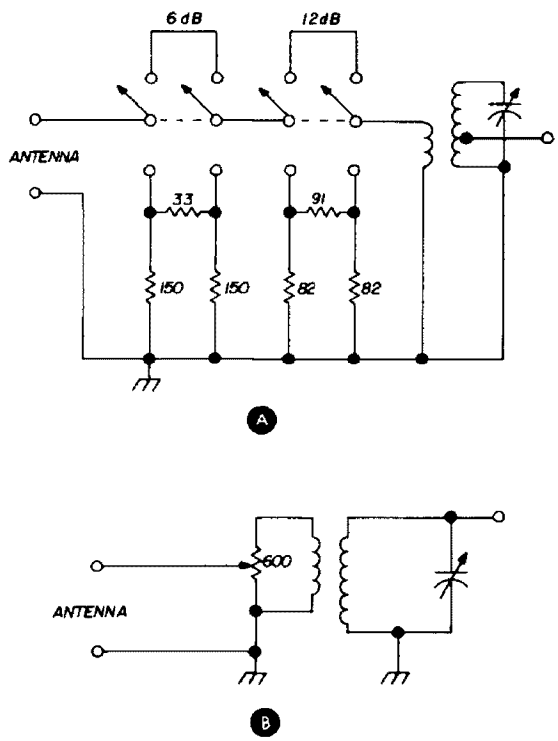


fig. 12. Two types of attenuators which may be used to control signal input to communications receivers. Step attenuator is shown in (A); simple potentiometer attenuator is shown in (B).

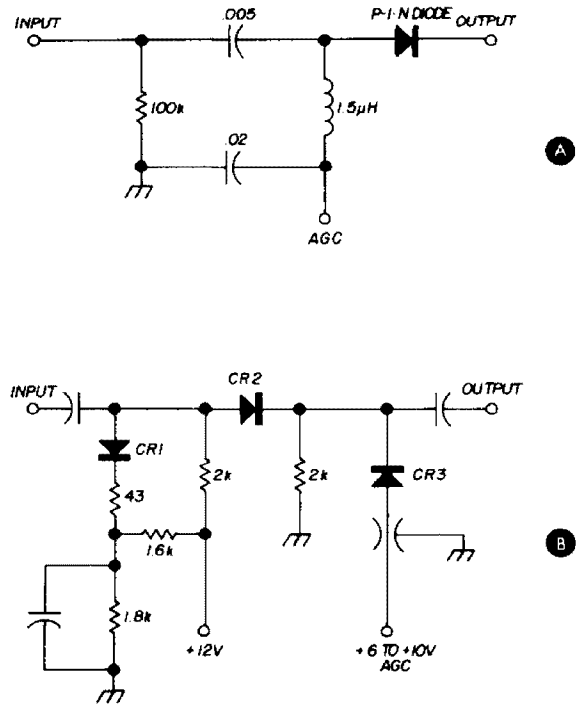


fig. 13. Agc voltage-controlled attenuator circuits using PIN diodes. In the circuit in (B), with +6 volts on the agc line, minimum attenuation is 1 to 2 dB (CR1 and CR3 off, CR2 on). With +10 volts on the agc line, mixmum attenuation is 38 dB (CR1 and CR3 on, CR2 off).

voltage-controlled attenuator using PIN diodes ahead of the first stage. A single diode attenuator is shown in **fig. 13A** and a more complex three-diode version with both forward and reverse biased diodes is shown in **fig. 13B.**[7,8] The designer must be alert to the possibility that the cross-modulation level of the diodes can be less than that of the amplifier or mixer device that they are intended to protect.

selectivity

Rf selectivity is important in the search for the ultimate in strong-signal capability since a perfect rf filter will eliminate all forms of strong signal interference. Crystal-lattice filters are available throughout the high-frequency range and are the near perfect answer for fixed frequency operation.[9,10] Unfortunately, if you are interested in more than a few fixed channels you must go on to less perfect filters.

Helical resonators are perhaps the next best rf filters (**fig. 14B**). They are not easily adaptable to continuous tuning and

they become quite bulky in the high-frequency range, but they can give Qs of 1000 or more which results in bandwidths of 1 kHz per MHz; 14 kHz at 14 MHz for example.[10,11,12,13]

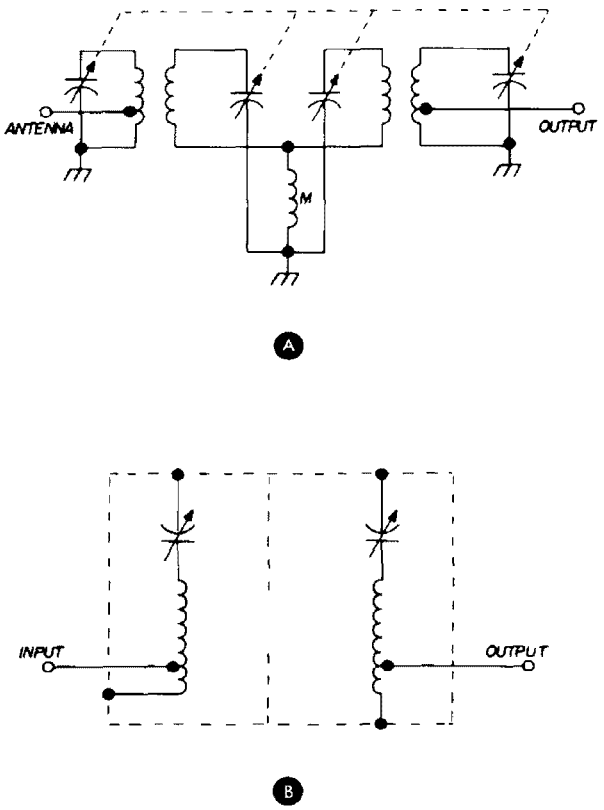


fig. 14. Four-circuit rf filter (A), and helical resonator (B).

Cascaded tuned circuits are also effective and most modern receivers have at least two such circuits between the antenna and first stage, and some have as many as four. One arrangement is shown in fig. **14A**.

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ham radio



integrated-circuit speech clipper

Rf clipping with the LM-373 i-f strip IC provides low distortion and superior bandpass shaping

Charles G. Bird, K6HTM, 875 Lindo Lane, Chico, California 95926

Almost all types of speech processors, and their vices and virtues, have been discussed in these pages. The "audio-rf-audio" low-level processor, as represented by the Comdel CSP-11, has yet to be examined in detail. The method used in the Comdel unit provides speech clipping and filtering analogous to that of rf clippers, but the processed waveform is supplied to the ssb exciter input stage at audio level. A significant increase in average-to-peak talk power is claimed.[1]

operating principles

A block diagram of the Comdel unit appears in fig. 1. A closed-loop system is used, with output at the original audio frequency. Instantaneous amplitude limiting occurs, but does not introduce any harmonic components.[1]

The microphone input signal is amplified and mixed with a local-oscillator signal, which is common to the balanced modulator and product detector. Two filter stages are used, which must be capable of passing a signal for ssb service. The filtered signal is clipped, passed through another (identical) filter and applied to a product detector. The resulting audio signal, which is further processed by a low-pass filter, may be applied to an ssb exciter input stage.

The unit to be described consists basically of two ICs, a fet local oscillator, a balanced modulator, and two mechanical filters.*

*A possible source for the filters is Star-Tronics, Box 17121, Portland, Oregon 97217. The MFC-8040 and LM-373 are available from Circuit Specialists, Box 3047, Scottsdale, Arizona 85257. Helpful information on the MFC-8040 and LM-373 will be found in the bibliography at the end of the article. editor.

In fig. 2 a Motorola MFC-8040 is used as a low-noise audio stage, followed by a diode ring modulator and an usb 500-kHz Collins mechanical filter (BW = 2.75 kHz), which drives the agc stage of a National Semiconductor Products LM-373. Output from this stage at pin 9 is clipped and filtered by a second filter and reinserted at pin 4 of the LM-373. causes the MFC-8040 to overload on loud speech. The input to this stage is therefore shunted with a level pot. A subminiature Allen-Bradley unit is available in a blister pack, which can be easily mounted on the circuit board.

The modulator and clipper diodes were on hand and were carefully matched for low forward resistance. There may be

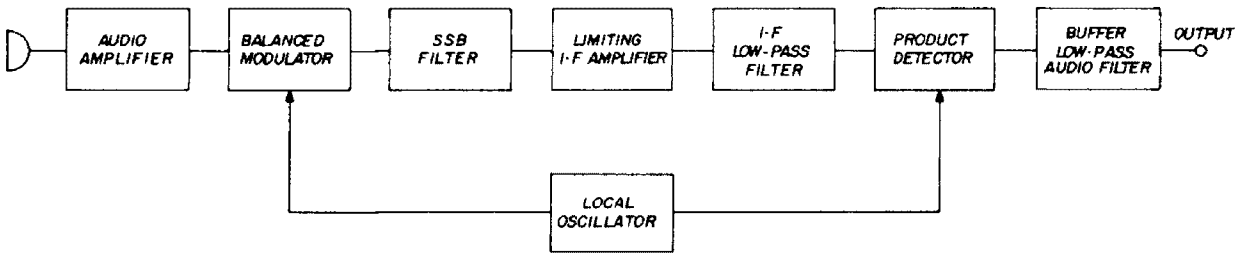


fig. 1. Block diagram of the Comdel CSP-11.

Local-oscillator output at 500 kHz is coupled to the balanced modulator and also to pin 6 of the LM-373; audio is derived from pin 7.

The beauty of the LM-373 is that its upper frequency limit is 15 MHz, thereby permitting the use of any popular filter, whether home-made with low-frequency surplus crystals or with readily available 9-MHz units.

circuit description

The schematic of the IC processor is shown in fig. 3. A microphone such as the Turner 454-C has excessive drive, which an advantage to using hot-carrier diodes in the clipper circuit.[2]

No special attention was paid to obtaining carrier suppression other than observing physical symmetry in layout of modulator components, since the carrier is common to both modulator and detector. If it's desired to use the unit as an exciter base by providing the processed signal after the second filter to an external connector, then greater precision in balancing would be required. Capacitor C1 was found to be about 15 pF. The rf output may be monitored at J1 with an oscilloscope or low-frequency receiver.

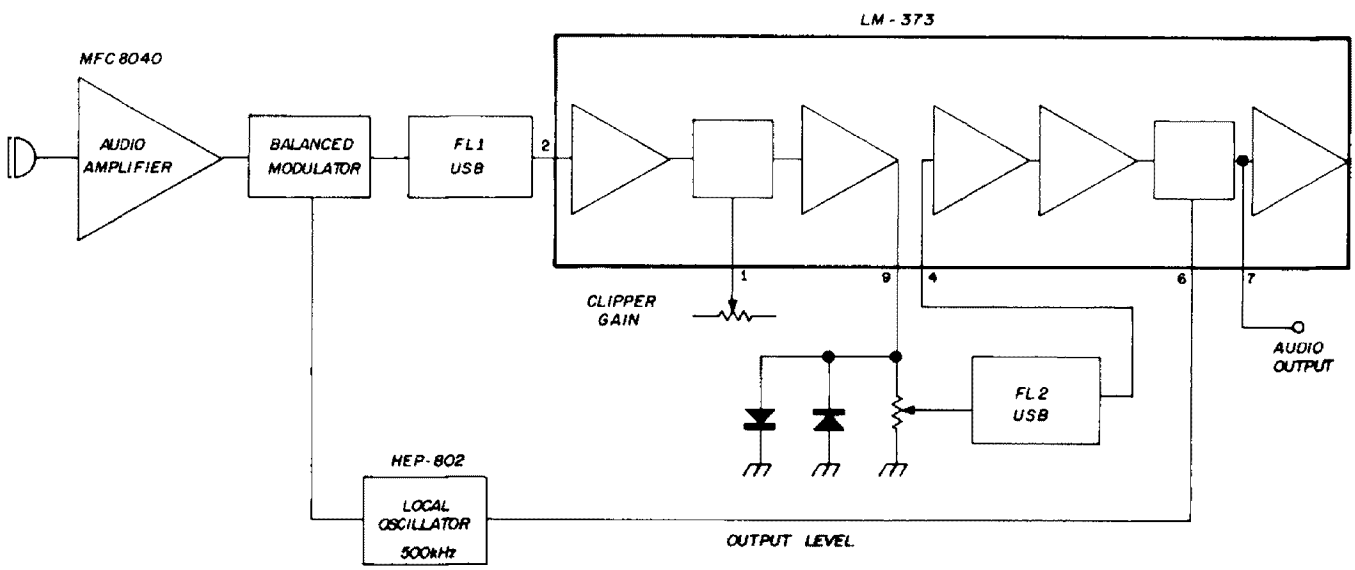
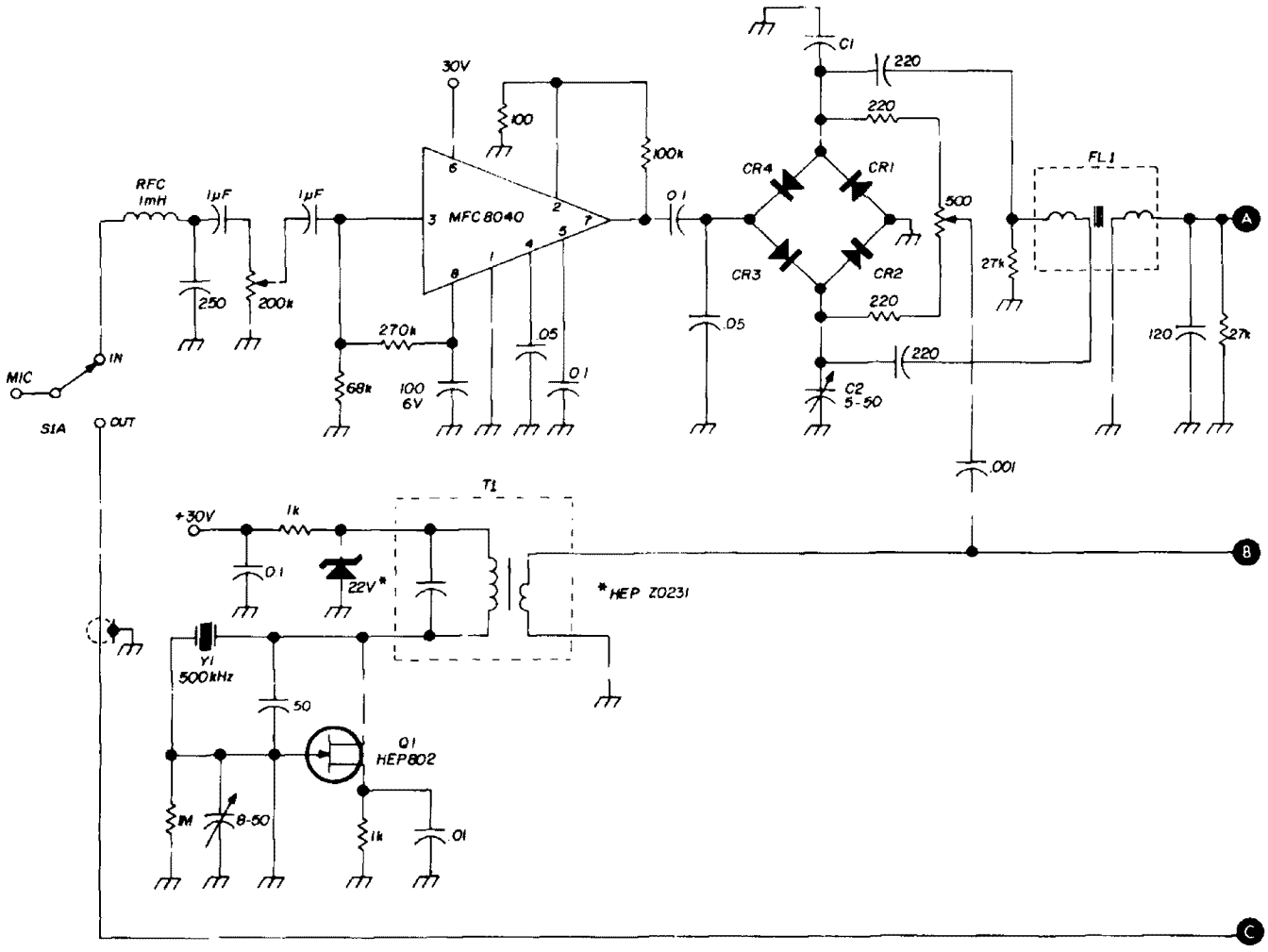


fig. 2. Block diagram of author's speech clipper using two ICs.



C1	value chosen to balance with C2
CR1-CR4	1N75 matched for low forward resistance
CR5,CR6	1N75, matched for low forward resistance, or hot-carrier diodes
FL1,FL2	Collins F500Z-12-6618 mechanical filter, 2.75 kHz passband
T1	455-kHz i-f transformer (Miller 8810I). See text
Y1	500 kHz. Channel 70 FT-241 holder, marked 27.0 MHz (JAN Crystals)

fig. 3. IC speech clipper schematic.

The output-level control should be fully open for best indication of carrier balance.

carrier oscillator

The 500-kHz oscillator uses a 455-kHz subminiature i-f transformer with the slug approximately half way out of the coil form. Oscillation will cease at a point in this direction, so adjustment is fairly critical. About 9 volts of rf can be measured on the primary and about 0.5 volt on the secondary. A 22-volt zener is used to regulate the voltage and protect the HEP-802.

power supply

The point at which the clipper gain control is set is quite dependent on the voltage applied to the LM-373. The load on a 12-volt battery, for example, will

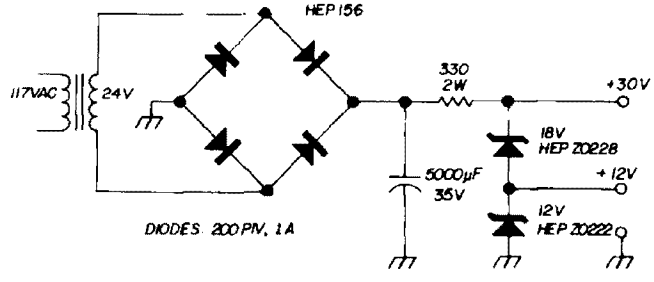
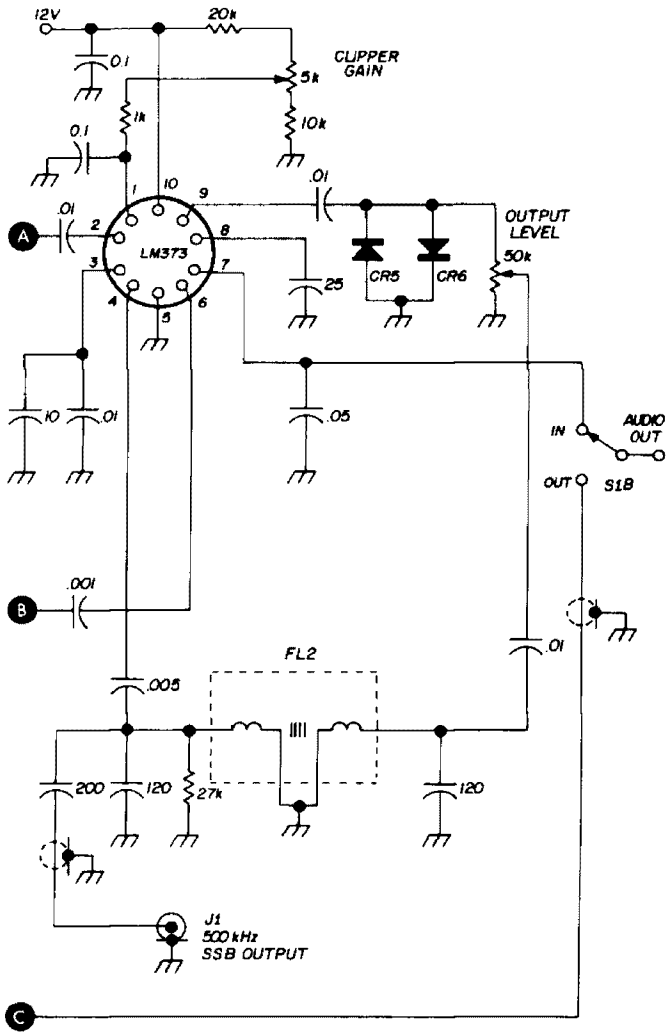


fig. 4. Power supply for the speech clipper.



shift the clipping point counter clockwise because of voltage drop. By the same token, a 30-volt battery used for the MFC-8040 and Q1 will be quickly depleted. A suitable power supply is shown in fig. 4.

adjustments

Preliminary adjustments are made to ensure that Q1 oscillates. (This circuit and helpful hints are given by Hank Olson, W6GXN. See reference 3.)

With the output connected to an audio amplifier, the operation of the LM-373 should be checked by listening for ambient noise. If voltage is applied to the MFC-8040 but no speech is heard, you can at least be assured that the LM-373 is working if the transient generated by the make and break voltage to the MFC-8040 is heard through the remainder of the unit. The problem can be isolated at the MFC-8040 or before. An audio oscillator is useful here. Rf filtering at the processor input is necessary.

The unit described is mounted on a 2-1/2 x 4-1/4-inch pc board and placed inside a 3 x 5 x 7-inch chassis turned on its side, with a hand-fashioned one-piece cowl placed over the sides and top. It is painted to match Drake styling.

conclusion

Performance was all that was hoped for, but as expected, not much better than that of a typical agc-type compressor in terms of additional sock to the signal. Direct comparisons were made with the speech processor described by K6PHT.[4] Scope displays were perhaps slightly fuller with the IC clipper described here. On-the-air reports seemed to favor the clipper for over-all quality, but did not necessarily indicate its superiority in increased average-to-peak signal strength.

It appears that low-level rf clipping is beneficial mainly with regard to low distortion and superior bandpass shaping. Another possible advantage over the agc-type compressor appears to be greater dynamic range; that is, a lower threshold of background noise for the same amount of increased average-to-peak power. Judicious adjustment of the clipper gain and output-level controls bears out this contention.

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vhf receiver scanner

Automatic channel monitor using TTL logic

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The purpose of a scanner is to allow several vhf receiving channels to be monitored automatically, thus eliminating manual switching between channels. In operation, a slow-running oscillator is turned on and off by the receiver squelch circuit. Flip-flops, controlled by the slow-running oscillator, alternately turn one receiver oscillator on and off, and the process repeats at the next receiver oscillator. When the scanned channel is active, the receiver squelch circuit stops the slow-running oscillator, which conditions the flip-flop so that the station can be monitored.

This article describes a scanner using TTL logic. Some of its features are summarized below.

1. Either positive or negative logic can be used for squelch control.
2. Oscillator switching and channel indication can be accomplished in several different ways.
3. Etched circuit board or breadboard layout can be used.
4. As many as 20 channels can be switched with the addition of compatible devices.

design considerations

It was desired to produce a versatile, inexpensive, simple, and reliable unit. TTL logic provides easy interface with a wide variety of receivers, is reliable, and has recently enjoyed price reductions.

operation

Three two-input NAND gates are used as an oscillator whose frequency is determined by a 1k resistor and 300 μF capacitor (**fig. 1**). The fourth NAND gate connects the oscillator to a counter (four internally connected flip-flops). When the oscillator gate is off (LO input), the counter remains at its last count. When the oscillator gate is on (HI input), the counter proceeds with an upward consecutive count to eight. The counter is reset

automatically by a jumper from pin 11 to pins 2 and 3 and begins counting again from zero. This action continues until the squelch circuit gates the oscillator off. As the scanner counts, it sequentially turns on one receiver oscillator at a time by from being fed back to the scanner. It may also be desirable to bypass the scanner output for rf with a 0.001-μF capacitor on each output.

Note that the scanner is connected for receiving only. If the scanner is installed

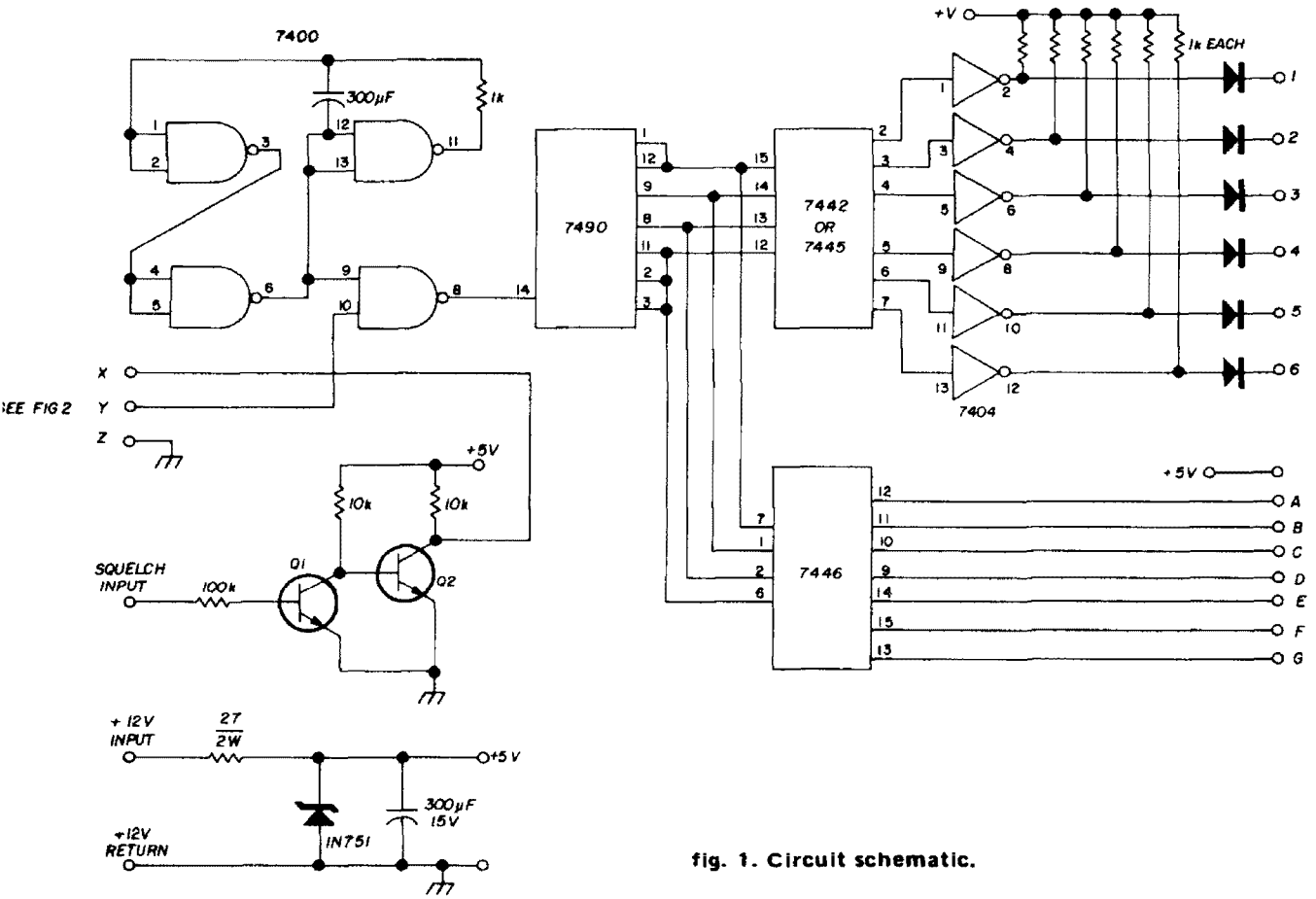
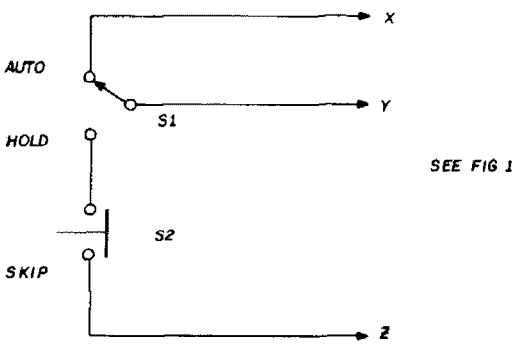


fig. 1. Circuit schematic.

either forward biasing a diode connected to a crystal, or by applying bias directly to the oscillator circuit. Either a seven-segment readout or individually marked lamps can be used as indicators to identify the channel being monitored. The seven-segment readout is driven by a special decoder-driver IC, which receives its input from the counter BCD output and translates the count to indicate the channel. Lamps may be driven directly from the decimal decoder. The oscillators can be driven from the decimal decoder directly or by an inverter following the decoder. depending on the voltage sense required to switch your receiver oscillator circuit. It is recommended that series diodes be used at the scanner output to the oscillators to prevent receiver voltages in a transceiver that uses a common switch to change both receive and transmit channels, ensure that the transmit crystal does not become enabled by the scanner, or you may transmit on several



S1 = SPDT TOGGLE
S2 = SPST NORMALLY CLOSED PUSHBUTTON

fig. 2. Suggested circuit for manual override control.

channels at the scan rate. It's best to choose a switch position for receive without crystals in either the receive or transmit socket. Use this switch position to enable the scanner. When in radio contact, switch the channel selector to the appropriate position and disable the scanner; that is, don't put the scanner in hold and attempt to transmit!

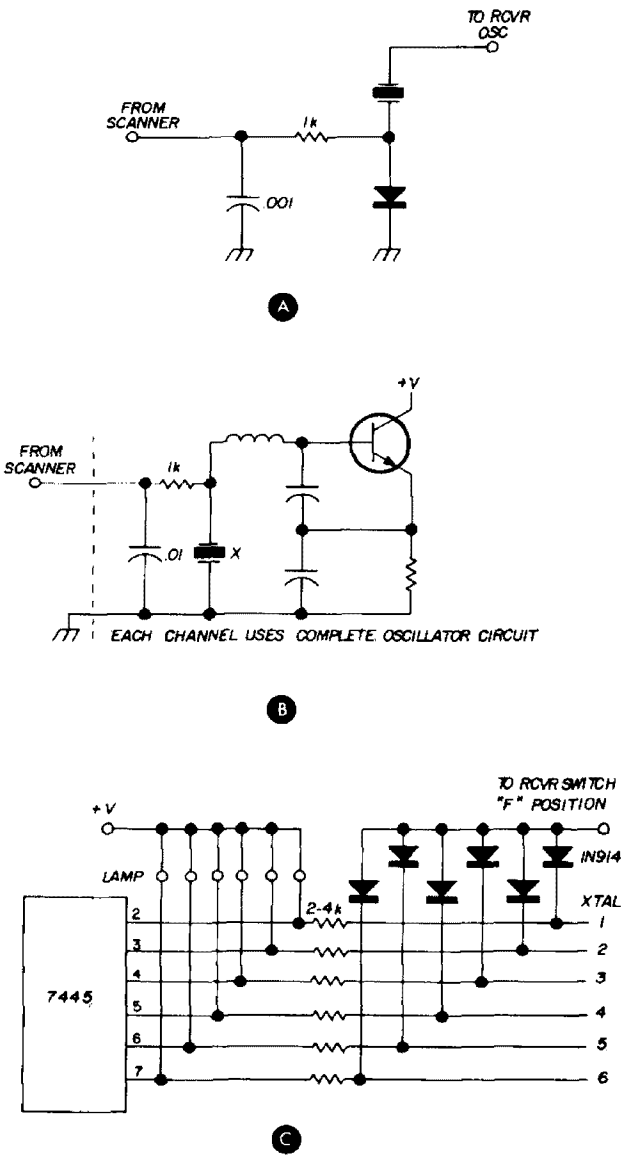


fig. 3. Receiver oscillator control circuits. Method A switches crystal directly; method B switches bias to oscillator. Circuit in C is for oscillator control in Regency receivers.

inputs

power supply. The unit requires +12 V at 220 mA if you use the zener circuit.

squelch. Provision is made to accept either:

A. A +5-volt signal to enable scan and a zero-volt signal or ground to stop scan.

B. A +5-volt signal to stop scan and a zero-volt signal or ground to enable scan.

The choice of A or B depends on the voltage available from your receiver squelch circuit. For situation A, use the complete input circuit with two transistors (**fig. 1**). For situation B, use only one transistor and be sure to use a high-value base resistor (100k or more) to prevent loading your receiver squelch circuit.

The scanner will work without the switches, but for convenience and flexibility, switches are recommended. This is especially true if one of the monitored channels is a repeater. A toggle switch can be used to change from auto to hold, and a momentary pushbutton switch can be used to skip from one channel to another. See **fig. 2.**

outputs

channel indication. If you are using a seven-segment readout, wire the unit as indicated for that section. If you are using indicator lights, one end of each lamp can be tied together and connected to a positive voltage, depending on the lamps used. The other end of each lamp can then be connected to the appropriate 7445 or 7442 terminal. The 7445 can handle up to 30 volts and sink as much as 80 mA, although low-current, low-voltage lamps are preferred.

control of oscillators

This scanner can switch oscillators and indicate channels in several different ways. The standard version, shown in **fig. 1**, is used with a seven-segment readout and controls an oscillator similar to that indicated in **figs. 3A and B** by applying forward bias to a diode or by biasing the oscillator directly. If more than 4.5 volts are needed to operate an oscillator, use an open collector inverter (7406) instead of the "totem pole" output inverter (7404). The 7404 should not be operated above 5 volts. The 7406 can be operated with 30 volts on the open output (although package V_{CC} must remain at +5 volts). If you

use lamp indicators, you can connect them to each 7442 output, provided the lamps require 5 volts or less.

special case — the Regency receiver

Some receiver oscillators don't lend themselves to an easy switching method, such as described above. Fig. 3C is an example of a scheme for switching oscillators in the Regency receiver. The diodes are at +8 volts and rf ground. Each diode cathode is connected to an oscillator crystal. The common side of the lamp is tied to +12 volts. When the 7445 ouptut is off, 12 volts appear across the lamp and resistor. This causes the diode to be reverse biased, keeping the crystal above rf ground. When the 7445 ouptut goes LO (ground) the lamp turns on, current flows from the +8-volt source through the diode and resistor to ground, which causes the diode to be forward biased. This action provides an rf path for the crystal to turn on.

Referring to the layout diagram, fig. 4, the following procedures also apply if the scanner is to be used with the Regency.

Insert diodes in place of the pull-up resistors. Insert resistors (2 to 4k; value determined by trial) in the 7404 socket between pins 1&2, 3&4, 5&6, 8&9, 10&11, and 12&13. (Do not use the 7404.) If you wish to use a digital readout, replace the lamps with 1k resistors and connect the common side to +12 volts.

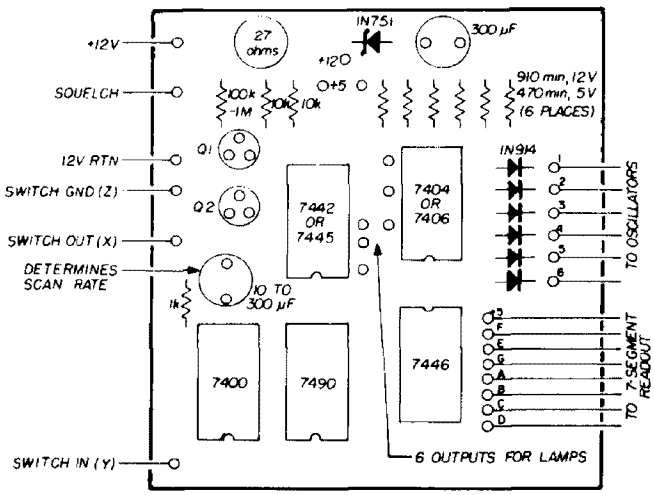


fig. 4. Component layout. Indicators can be numbered lamps or 7-segment readout devices.

construction

Assembly is straightforward. If an etched circuit board is used, follow the parts layout diagram, fig. 4. If you're breadboarding, any logical layout will work. Note proper device orientation: diode cathodes have a band: IC dot locates pin one. Inputs and outputs are labeled on the drawing. A parts list appears in table 1.*

table 1. Parts list.

item	description	quantity
SN7400	gate	1
SN7490	counter	1
SN7442	decoder	1
SN7446	decoder	1
SN7404	inverter	1
2N4140	NPN transistor	2
1N914	diode	6
1N751A	zener	1
300 µF	capacitor	2
27 ohm 2W	resistor	1
100k	resistor	1
10k	resistor	2
1k	resistor	7
7 segment readout		1
readout socket		1
board		1

summary

These examples indicate the versatility possible with the scanner. You can supply it with either positive or negative logic for squelch control. You can control either positive or negative logic oscillators. Higher voltage loads can be switched using the open-collector gates. A choice of lamps or seven-segment readout is available. Adding another inverter from the decoder and removing the counter reset allows ten oscillators to be controlled. By adding one flip-flop and another decoder, as many as twenty oscillators can be controlled.

ham radio

*This scanner is available in kit form from Hamtronics in Rochester, New York. For information, send a self-addressed, stamped envelope to Jerry Vogt, WA2GCF, Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612.

how to use Plessey SL600-series integrated circuits in amateur communications equipment

A complete description of the versatile Plessey SL600 linear ICs, and how to use them in receivers and transmitters

James M. Bryant, G8FNT, Plessey Linear Applications Engineer

The Plessey SL600 series of integrated circuits for use in transmitters, receivers and other communications systems is now in use all over the world, and Plessey continually receives inquiries from radio amateurs who want applications data on the different SL600 circuits. In the hope of satisfying this need, G8FNT, a linear applications engineer for Plessey, has prepared this article.

The Plessey SL600* series of ICs includes rf and i-f amplifiers with low cross-modulation and good agc; audio amplifiers with and without agc, high-performance balanced modulators and speech agc generators. Also included in the series is a complex circuit containing a-m and ssb detectors, and a CW-operated agc system.

This article describes some transmitters and receivers that are based on the use of SL600 ICs, but does not cover the audio sections or high-power rf amplifiers. The first part of the article describes a variety of communications systems which use Plessey ICs. The latter part of the article gives circuit details and comments on some potential causes of trouble experimenters might run into.

*The complete line of Plessey SL600 ICs is available from Circuit Specialists, Box 3047, Scottsdale, Arizona 85257. Write to them for their complete price list, or see their advertisement in this issue.

receiver systems

The simplest receiver which can be built from SL600 ICs is the direct-conversion unit shown in block form in **fig. 1.** This direct-conversion receiver, also known as a synchrodyne, can be used for the reception of a-m, ssb and CW. The vfo is tuned to the carrier frequency in the case of a-m and ssb, and a few hundred hertz away for CW. Upper and lower sidebands are equally well detected by this receiver, and if the audio passband is limited, it is quite selective.

However, if this simple direct-conversion receiver is used to receive an upper-sideband ssb signal adjacent to another ssb signal only 2 or 3 kHz away, the interference will be quite severe. This interference can be removed, and only one sideband detected, with the more complex phasing system shown in **fig. 2.**

The direct-conversion receiver in **fig. 2** uses rf and audio phasing to cancel one sideband, so it is truly a single-sideband receiver. In this circuit it is necessary to have accurate phasing of the signals, and well matched gain in the two audio channels feeding the summing stage. The upper or lower sideband may be selected by reversing the audio phasing. (The rf phase may also be reversed for the same purpose, but switching the audio is easier.) The system in **fig. 2** detects the lower side-band when the upper channel phase shift is positive.

Another direct-conversion receiver is shown in **fig. 3.** Whereas the circuit of **fig. 1** is subject to interference, is not very sensitive and has no agc, the circuit of **fig. 3** has rf filters to minimize cross-modulation, an rf amplifier, agc and perhaps, an S-meter. Depending upon the sensitivity required and the audio gain available, one or two rf amplifiers may be used. The SL610 IC has a gain of 20 dB, and frequency response to at least 146 MHz. (This performance, which exceeds that shown on the SL610 data sheet, depends on very careful layout, very short leads and very great attention to coupling and decoupling of power supplies and agc. However, amateurs who use the SL610 on two meters find that the performance at these frequencies is satisfactory.)

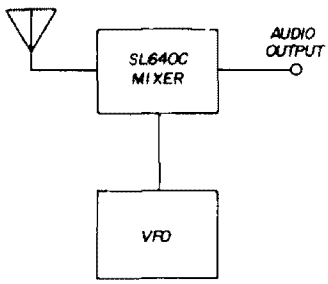


fig. 1. Block diagram of the most simple direct-conversion receiver which can be built with Plessey SL600 ICs.

Other ICs which may be used for the rf amplifiers are the SL611 and SL612. The SL611 provides 26 dB gain and is usable to 80 MHz; the SL612 has 34 dB gain with frequency response to 15 MHz. The amplifier you choose, here and in the other systems to be described, depends upon the frequency and the gain required. The SL612 has the extra advantage of lower current consumption and slightly lower noise figure.

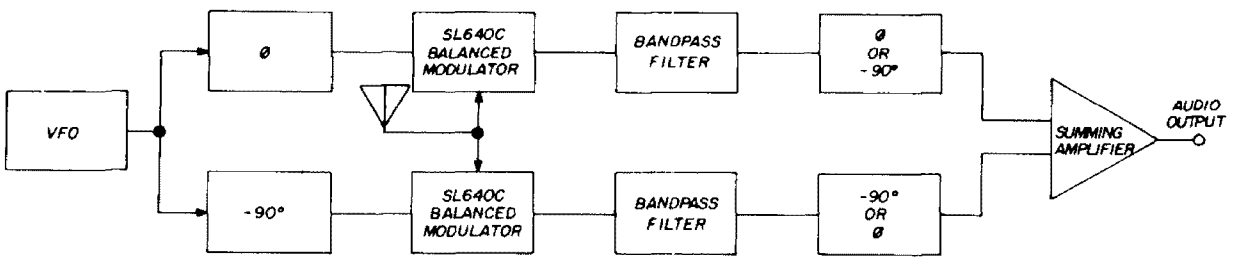


fig. 2. Block diagram of a phasing-type direct-conversion ssb receiver.

conventional superhet

A much more conventional superhet receiver is shown in **fig. 4.** This design consists of an rf stage with agc (which

would probably be an SL610 IC), a mixer (SL640 or SL641), an i-f filter (LC, crystal or ceramic), an i-f amplifier with agc and a detector.

The i-f amplifier could be one or two stages, depending on the sensitivity required, but agc would normally be applied to only one stage. An SL640 or SL641 would be suitable for the ssb and CW detector, followed by an SL621 to provide audio agc. If a-m operation is required, an SL623 could be used. This IC also generates carrier agc, and when used with a bfo, detects CW or ssb.

For fm detection either the SL432 or SAA570 may be used, but a separate carrier detector is required to provide automatic gain control.

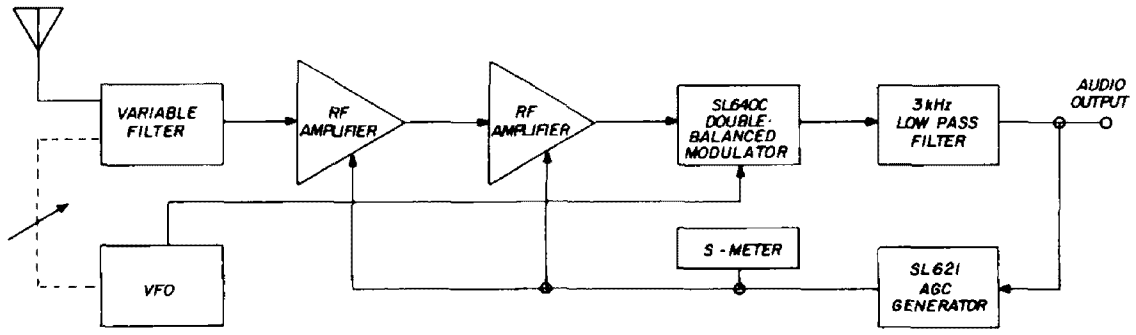


fig. 3. Practical direct conversion receiver can be used successfully on the amateur high-frequency bands.

agc output of SL623 and the audio-derived agc provided by an SL621 connected to the ssb output of the SL623.

The i-f filter may also be switched, with a narrow bandwidth filter for ssb, and a wider one for a-m. To detect narrowband fm, a detector such as the SL432 or SAA570 is connected to the output of the second SL612C. During narrowband fm reception agc is taken from the SL623.

Double-conversion superhets may also be designed using SL600-series ICs, but with modern filters, double-conversion superhets are rarely needed on the amateur high-frequency bands. However, they are occasionally used on uhf or where complex tuning systems are required.

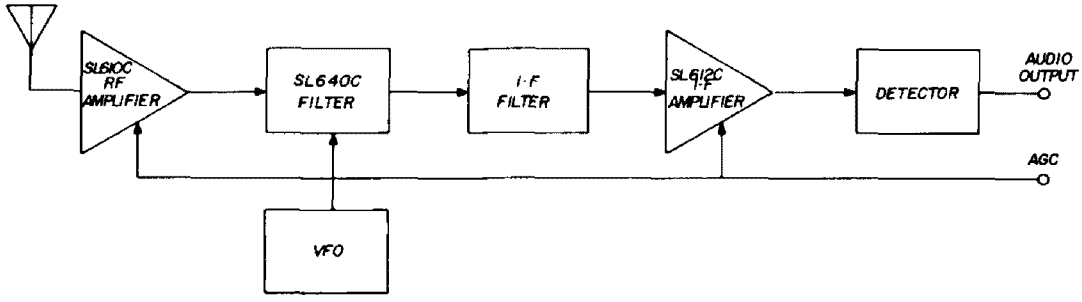


fig. 4. Layout of the basic superhet receiver.

A more complete superhet circuit which features front-end tuning and both a-m and ssb detection is shown in **fig. 5.** The output of the audio stage is switched between the two detector outputs of the SL623 IC. The agc line, which also drives an S-meter, is switched between the CW

Inasmuch as the same techniques are used as in the single-conversion superhets shown in **fig. 4** and **fig. 5,** no double-conversion designs will be described here. However, it should be noted that SL600 devices have high gains, and you should avoid using too many amplifying stages.

ssb exciters

A basic filter-type ssb exciter is shown in **fig. 6.** The audio and a low-frequency rf signal from an oscillator (the bfo if the exciter is part of a transceiver) are mixed in an SL640 IC. The SL640, as a result of

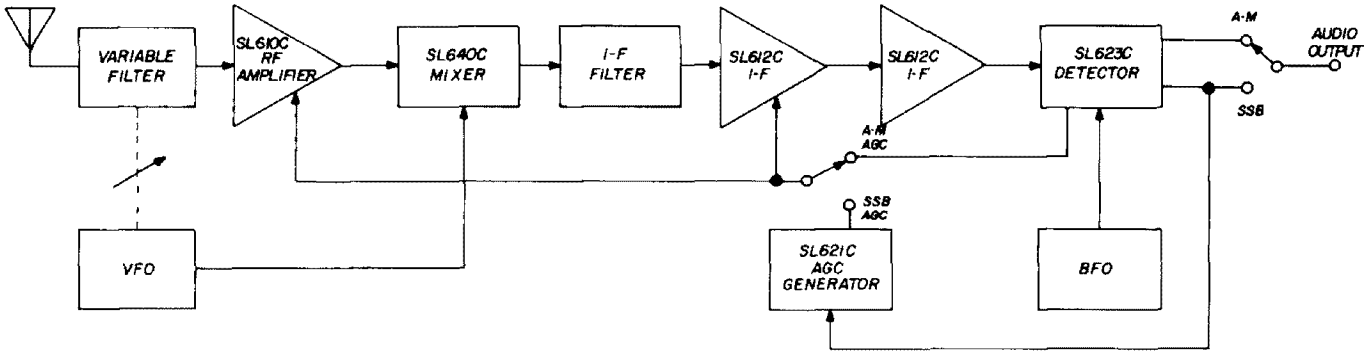


fig. 5. Complete superhet communications receiver for ssb, a-m and CW. Fm detection can also be added as described in the text.

its good carrier rejection, provides a clean dsb suppressed-carrier signal at the output. This dsb signal is passed through a narrow bandpass filter to remove one sideband. The remaining sideband is converted to the final operating frequency by another SL640. The image is removed by a filter, and the output is connected to a linear amplifier.

A more complete filter type ssb exciter with automatic level control (alc) is shown in **fig. 7.** In this circuit the

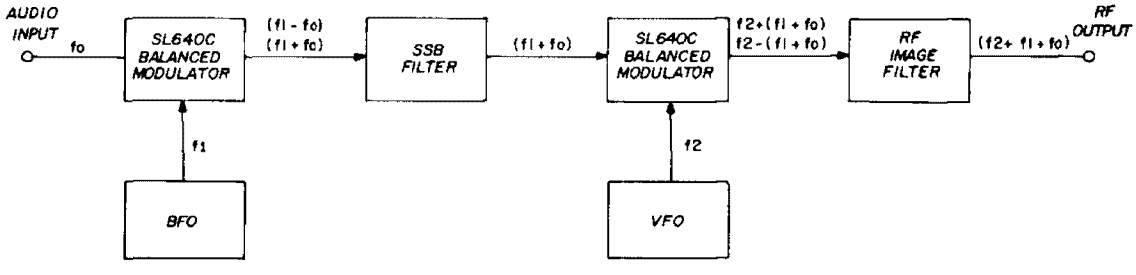


fig. 6. Basic filter-type ssb exciter.

SL610C amplifier is controlled by an alc signal which, in most cases, is derived from the final amplifier stage, either by a threshold detection system or by grid-current detection in the output tube.

A phasing-type ssb exciter is shown in **fig. 8.** The audio speech signal, which must normally have limited bandwidth, is phase shifted so that two separate, equal-amplitude audio signals with a 90° phase difference are obtained. These audio signals are applied to the signal inputs of two SL640 ICs, rf reference and quadrature signals are applied to the carrier inputs, and the two output signals are summed.

If the audio and rf reference signals are applied to one modulator, and the quadrature audio and rf signals are applied to the other, the lower-sideband outputs will be in phase and will add. The upper-sideband outputs will cancel, providing a lower-sideband output. Similarly, if the audio reference and carrier quadrature signals are applied to one modulator, and the audio quadrature and carrier reference are applied to the other, upper-sideband output is obtained.

This method of generating ssb appears attractive in many respects, and has the advantage that no expensive filters are required. Also, the carrier frequency may be varied so no further frequency conversion is necessary. Unfortunately, the

phasing systems has one serious drawback: To keep the undesired sideband 40 dB below the desired sideband, the phasing, both audio and rf, must be accurate within 2°. Also, the amplitude of the rf carrier applied to one modulator must be adjusted to minimize second sideband

figs. 5 and 7 or figs. 2 and 8, you'll note that ssb transmitters and receivers of the same types are quite similar. Therefore, with a little signal switching, it is possible to make one set of SL600 ICs perform as both a transmitter and a receiver. This, of course, saves both on SL600 ICs and

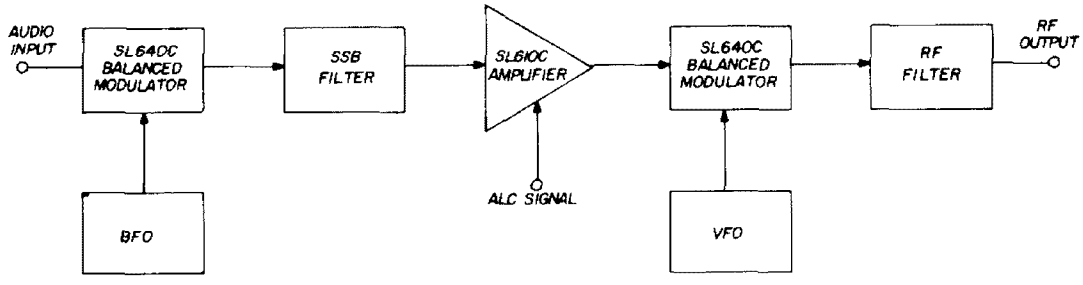


fig. 7. Ssb transmitter with automatic level control (alc).

generation, and carrier leak must be minimized in both modulators.

However, despite the adjustment problems, this method of obtaining a ssb signal is very popular, probably because no expensive filters are required. The system shown here is compatible with the direct-conversion receiver in fig. 2, and a very simple ssb transceiver can be built by combining these two systems.

amplitude modulation

Since a-m is merely double sideband with carrier, an SL640 may be used as an amplitude modulator if carrier leak is increased. If a 15k resistor is connected between pin 2 of an SL640 and ground as shown in fig. 9, there will be sufficient carrier on the output of the SL640 for a-m. By switching the resistor in and out of the circuit either a-m or dsb may be produced. If you also switch the filters following the SL640, a-m, dsb or ssb may be obtained from the same SL640 with the same inputs. With this simple arrangement you can build a multi-mode transmitter with very few components.

transceivers

If you look over the illustrations in

*Vogad is an acronym for voice-operated gain-adjusting device.

expensive filters. Fig. 10 shows the block diagram of a typical ssb transceiver based on the use of SL600-series ICs.

rf clipping

As has been noted by a number of designers, audio limiting and audio clipping are not useful techniques for increasing the average-to-peak power ratio of ssb transmitters, although audio agc is (derived, perhaps, from an SL622 vogad* circuit). If clipping is used, it must be performed on the rf ssb signal and the clipped signal must be filtered to remove splatter.

The rf clipping system shown in fig. 11 needs careful initial adjustment but yields remarkably good results. The input audio, which should be agc controlled, is con-

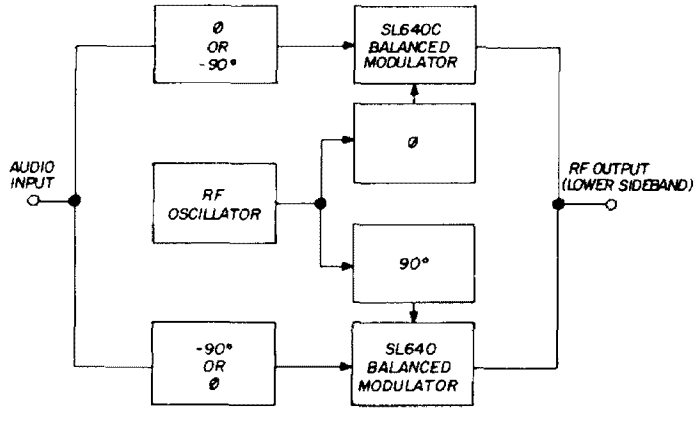


fig. 8. Phasing-type ssb exciter.

verted to ssb in the basic system; then the ssb signal is clipped by a symmetrical peak clipper. The signal is then refiltered to remove splatter and is passed through an alc amplifier and converted to the final operation frequency. The audio input level must be adjusted so that clipping is not so excessive that it destroys intelligibility.

If the clipper is replaced by a Schmitt trigger, and the audio input is given 12 dB per octave pre-emphasis above 1 kHz, a class-C power amplifier may be used. The output signal is received as ssb, although it has slight distortion. This arrangement provides a peak power equal to mean power during speech. If carrier leak is allowed to occur during pauses in speech so that the transmitter is always delivering the same power to the antenna, tvi is much reduced. In this case alc — and hence the SL610C — is not needed.

other systems

The Plessey SL600 integrated circuits may also be used in various other parts of amateur communications equipment. **Fig. 12** shows a mixer vfo which mixes the output of a low-frequency vfo with a crystal-controlled signal to produce a stable high-frequency vfo. With this circuit in a multiband receiver, several crystals may be used to tune several different bands with one vfo.

An IC agc system which is designed to stabilize the amplitude of an rf carrier is shown in **fig. 13.** A simple low-power IC linear amplifier is shown in **fig. 14.** In this circuit the value of the emitter resistor depends upon the power transistor used in the output stage.

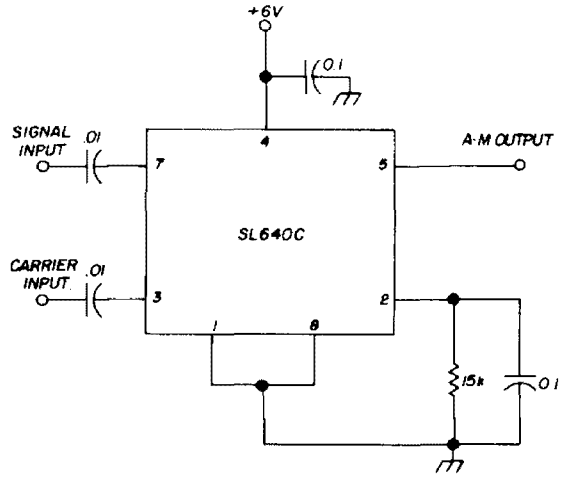


fig. 9. Amplitude modulator using a Plessey SL640 IC.

If audio squelch is required in a receiver, the system shown in **fig. 15** will provide it. This circuit is based on the fact that if pin 7 of an SL630 audio amplifier is grounded, the circuit is muted. The circuit illustrated, when turn-

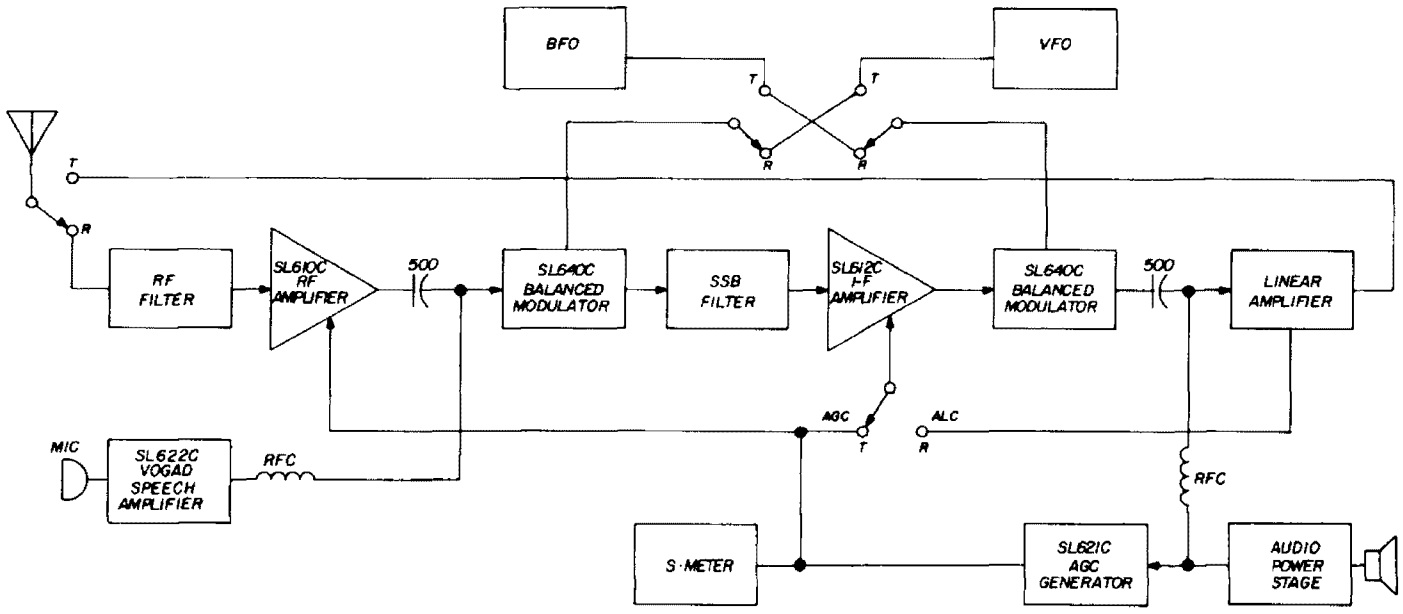


fig. 10. Complete ssb transceiver based on the use of Plessey SL600 ICs. During transmit the SL610C, SL621C and audio output stage are turned off; during receive the linear amplifier and SL622 are turned off.

ed on, insures that the SL630 is muted until the agc reaches a preset level. This prevents unwanted receiver noise when no signal is being received. The agc signal may be derived from an SL621, SL623 or

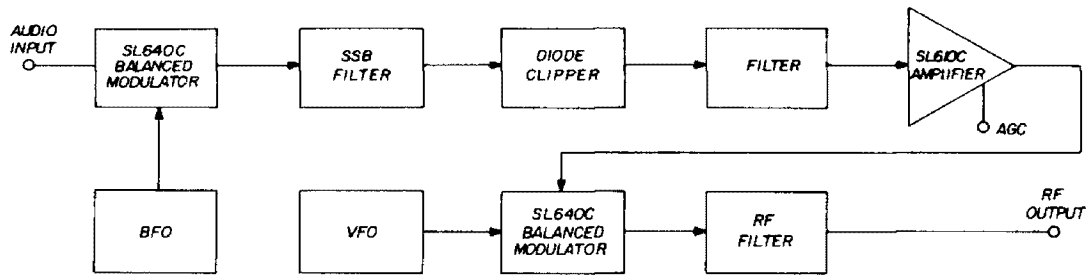


fig. 11. Ssb transmitter with rf clipping.

other sources. Any high-beta silicon NPN transistors are suitable for this circuit — a Plessey SL301C monolithic dual transistor is shown here.

A VOX system may be easily added to any transceiver which uses the SL622 as a microphone amplifier. A possible circuit

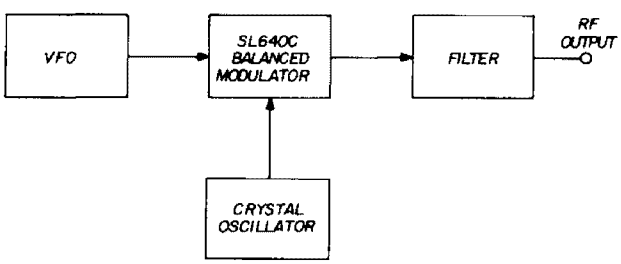


fig. 12. Basic circuit for the mixer-vfo.

is shown in fig. 16. It consists of an op-amp which is switched by the agc voltage of the SL622, and in turn switches the transmit-receive relay in the transceiver. The transistor may be any high-gain silicon type which can carry the relay current. However, a Darlington arrangement must be used to make sure the relay turns off again. This is because the minimum output of the op-amp can sometimes be more than 0.7-volt above the negative supply line; this is sufficient to turn on a single transistor.

communications receiver

The communications receiver circuit shown in fig. 17 is the same circuit shown in block form in fig. 5. Perhaps the easiest way of explaining the use of the SL600 family of ICs is to describe this circuit and its operation in detail.

Starting from the antenna, the input filter depends upon the i-f and the band being tuned. It must be sufficiently narrow to provide rejection at the image frequency (that frequency on the other side of the local oscillator from the wanted frequency and spaced the same amount from the oscillator frequency). If the image frequency gets through the input and mixes with the local oscillator it will produce an unwanted i-f signal.

The SL610 rf amplifier is coupled to the input filter in such a way that the IC is never inductively terminated. This is because the SL610 can become unstable if it looks into an inductance. The input coupling is also chosen so that the input filter Q is not lowered.

If an SL610 input is connected to a source which might be inductive, the source should either be shunted with a few thousand ohms, or a few hundred ohms should be connected in series with the input to the IC. The SL610 is biased by connecting the bias pin (pin 6) direct-

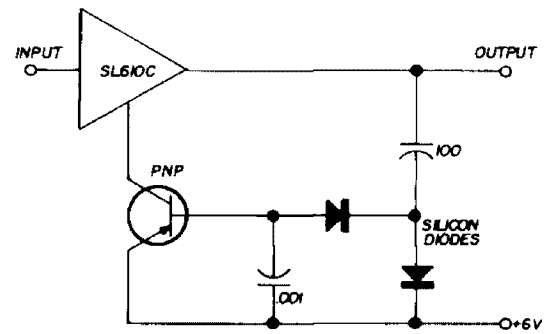


fig. 13. Carrier agc system uses one IC.

ly to the input (pin 5). (This same biasing arrangement is used with all the other rf and i-f amplifiers in this receiver.)

If an SL610, SL611 or SL612 are connected as shown in fig. 18, slightly

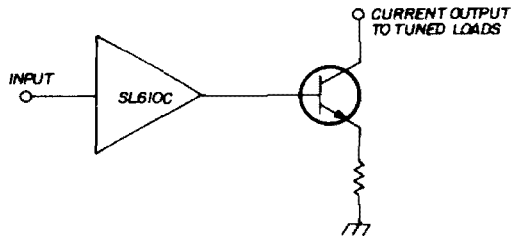


fig. 14. Low-power linear amplifier stage.

lower noise will result, but this is not usually worth the extra complexity. It is important that the input and output grounds on these devices be kept separate — output currents flowing through the input ground leads tend to produce instability.

Both the agc line and the SL610 positive supply (which is shared with the SL640 mixer) are decoupled to ground. Ideally, this is not necessary, but rf on power supply and agc lines can cause trouble with some circuit layouts. Where expense does not rule out decoupling, it is recommended. (The SL640 supply is not internally decoupled although that of the SL610 is.) To minimize the output current loop of the SL610 the SL640 ground (pin 8) should be as near as possible to the output ground of the SL610 (pin 8).

The output of the SL640C first mixer drives the input of the crystal filter. The

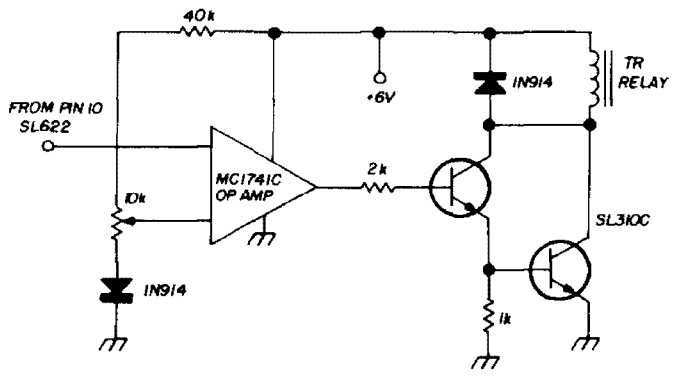


fig. 16. Simple VOX circuit for use with SL622C.

filter must be terminated by the correct impedance (pure resistance or resistance shunted by capacitance). If the resistive component is low enough, an SL641 may be used in the mixer circuit as shown in fig. 19. This is the case where SL640s are used: They may be replaced by SL641s in certain circumstances.

When the output from the SL640 is taken at pin 6, it requires an external load greater than 560 ohms. This is because the pin-6 output is a low-impedance emitter follower and it must not be used to drive capacitive loads. Some filters have transformer-type inputs with low dc resistance to ground (less than 10 ohms). In this case the terminating resistor, as long as it is larger than 560 ohms, may also be used as the load resistor. In this

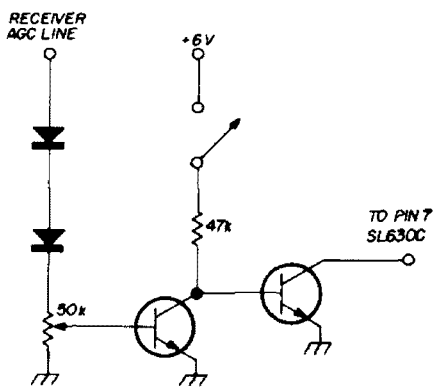


fig. 15. Audio squelch system.

case the dc blocking capacitor must *not* be used.

Pin 2 of the SL640 and SL641 must be decoupled to ground through a low-leakage capacitor (less than 0.1 μA). The reactance of this capacitor must be less than 10 ohms at the lowest input or carrier frequency. The carrier input from the local oscillator should be 100 to 200 mV rms and should be as free from modulation as possible.

The broadband i-f amplifier following the crystal filter consists of two SL612 ICs; agc is applied to only one stage at pin 7. The SL612 has a 70-dB agc range, while the agc range of the SL610C rf amplifier is 50 dB, giving a total for two gain-controlled stages of 120 dB. If both SL612 i-f amplifiers were agc controlled,

the total would be 190 dB, which is too much.

The positive power supply to the i-f stages and the SL640 detector is decoupled, and care must be taken to make sure that ground currents at the output of the i-f strip cannot flow near its input as this would lead to instability. The best grounding arrangement for the i-f strip is shown in fig. 20. No other connections should be made to the i-f strip ground.

The crystal filter must be properly terminated at its output. The input of an SL612 IC is approximately 5k in parallel with 4 pF. If necessary, this should be shunted (at ac only) by other resistors and capacitors to provide the correct filter-terminating impedance.

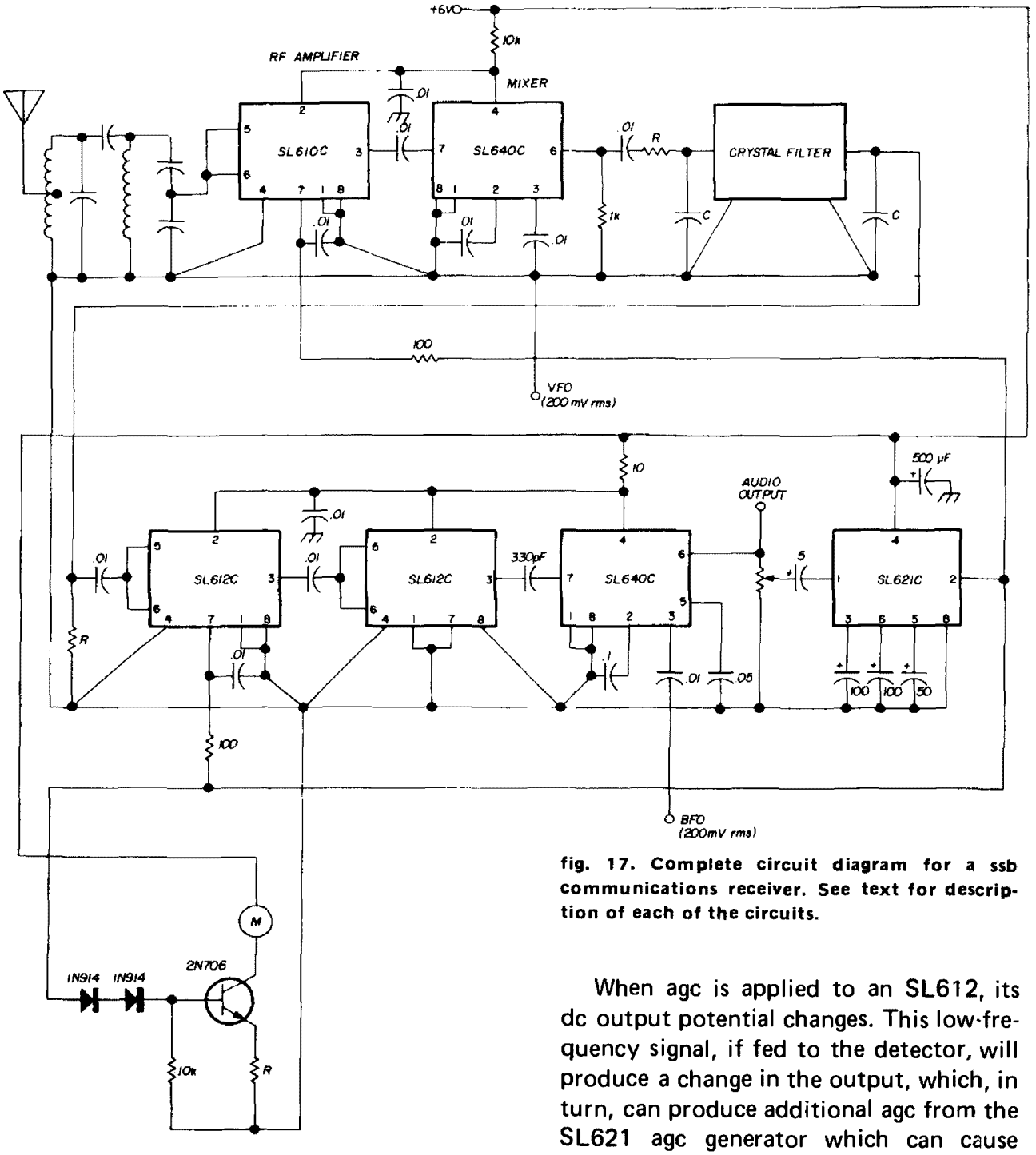


fig. 17. Complete circuit diagram for a ssb communications receiver. See text for description of each of the circuits.

When agc is applied to an SL612, its dc output potential changes. This low-frequency signal, if fed to the detector, will produce a change in the output, which, in turn, can produce additional agc from the SL621 agc generator which can cause very low-frequency stability or motorboating. To prevent this, the coupling capacitor between the last SL612 i-f stage and the SL640 detector should be as small as possible – about 330 pF is usual if the i-f is over 1 MHz. Another solution to this problem is to place a tuned circuit

at this point; this prevents the trouble and also reduces noise produced in the broadband i-f stages.

The output of the SL640 detector is decoupled to ground at frequencies above 4000 Hz with the 0.05 μF capacitor from pin 5 to ground. The load across pin 6 is a 1000-ohm preset potentiometer. The audio output to the amplifier is taken directly from pin 6, but the audio for the SL621 agc stage is taken from the potentiometer wiper. With this arrangement the agc threshold can be adjusted so that noise in the antenna and receiver circuits does not turn on the agc in the absence of a signal. The coupling capacitor to the SL621 should not exceed 1 μF or low-frequency instability may result.

The SL621 agc generator will usually drive a 500-μA S-meter connected in series with 5100 ohms and three silicon diodes from the agc output to ground. Occasionally, however, this is too much load for an SL621, and the transistor S-meter circuit shown in fig. 18 is preferable. The value of the emitter resistor depends upon the meter used, and is given by the following formula:

$$R = \frac{2.7}{I}$$

where I is the full-scale current in mA and R is in kilohms. For a 1-mA meter, this works out to be 2700 ohms. With this circuit the S-meter reads linearly in dB; from zero to full scale is about 120 dB.

The power supply to the SL621 must be well decoupled; 500 μF from pin 4 to ground is usually sufficient, but if the audio output stage shares the same power supply, this value should be increased. If a series-stabilized power supply is used, it should have a source impedance of less than 1 ohm.

The audio output stage may be an SL630, SL402, SL403 or other suitable audio power amplifier.

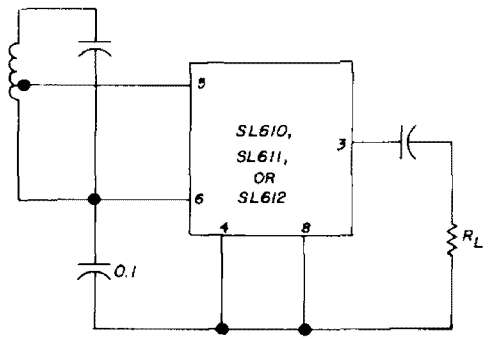


fig. 18. Low-noise signal coupling to the SL610 SL611 or SL612.

other applications

When SL600-series ICs are used in a transmitter or a transceiver they are used in much the same way as described above. However, since transmitters often contain rather large rf fields, particular attention must be paid to shielding and decoupling. In some cases it may be necessary to decouple individual stages.

When mixing frequencies or generating ssb in a transmitter, the original input frequencies are not wanted in the output. The SL640 and SL641 have approximately 30 dB signal and carrier rejection, but this may be increased with the circuit shown in fig. 21. With signal, but no rf carrier, R1 is adjusted for minimum carrier leak. All modulators used in transmitters may be adjusted in this way

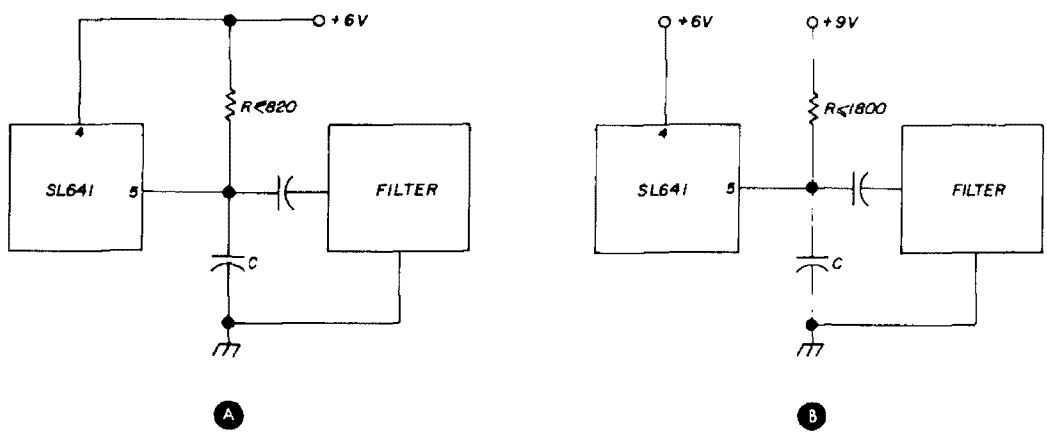


fig. 19. Circuits for matching the SL641 IC to filters.

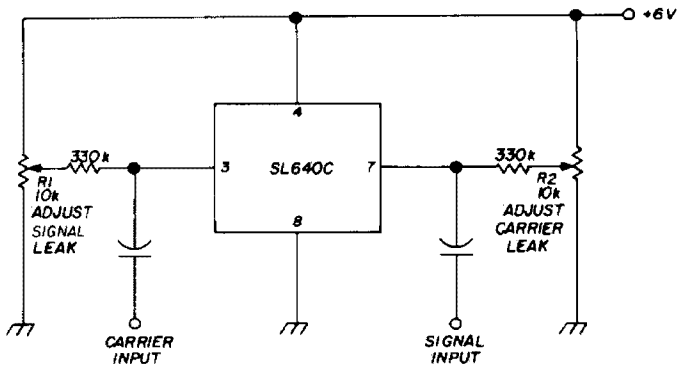


fig. 21. Signal and carrier-leak adjustment for the balanced modulator.

although it is less important in filter-type systems than in phasing systems.

The agc characteristics of the SL610, SL611 and SL612 are temperature dependent. It is unwise to use a voltage on an agc pin to set the gain of a stage. However, this may be done where agc is applied to another stage in the chain to compensate for signal variations.

The SL610, SL611 and SL612 tend to oscillate if they are required to drive a capacitive load. Such loads should be buffered by a resistance or another type of amplifier. For the SL610 and SL611 a 47-ohm buffer resistor should be used; a 150-ohm buffer resistor is used with the SL612. When an rf signal is taken from any of these SL600 amplifiers to points far removed, care must be taken to prevent instability caused by ground loops.

using the SL623

The SL623 IC combines the functions of a low-level, low-distortion a-m detector and agc generator with a ssb demodulator. It was designed specifically for use in ssb/a-m receivers, and is not used in ssb-only receivers. A typical application for the SL623 is illustrated in fig. **22.** Since the agc output from this circuit is CW derived, if audio-derived agc is required for ssb operation, an SL621 should be used. In the circuit of fig. **22** the resistor between pins 2 and 5 sets the value of carrier at which agc operation begins. The bfo signal should be a clean sine wave at about 100 mV rms.

When using the SL623 all the decoupling capacitors should go to one point. Also, the positive power supply should be decoupled. The SL623 is as sensitive to ssb as the SL640, but it requires 125 mV rms of a-m to activate the agc. Therefore, greater i-f gain may be necessary. Despite statements to the contrary in the data sheet, the SL623 functions to at least 30 MHz and, with reduced performance, to over 120 MHz.

other devices

The other members of the SL600 family are the SL620, SL622 and SL630. The SL630 is an audio amplifier with

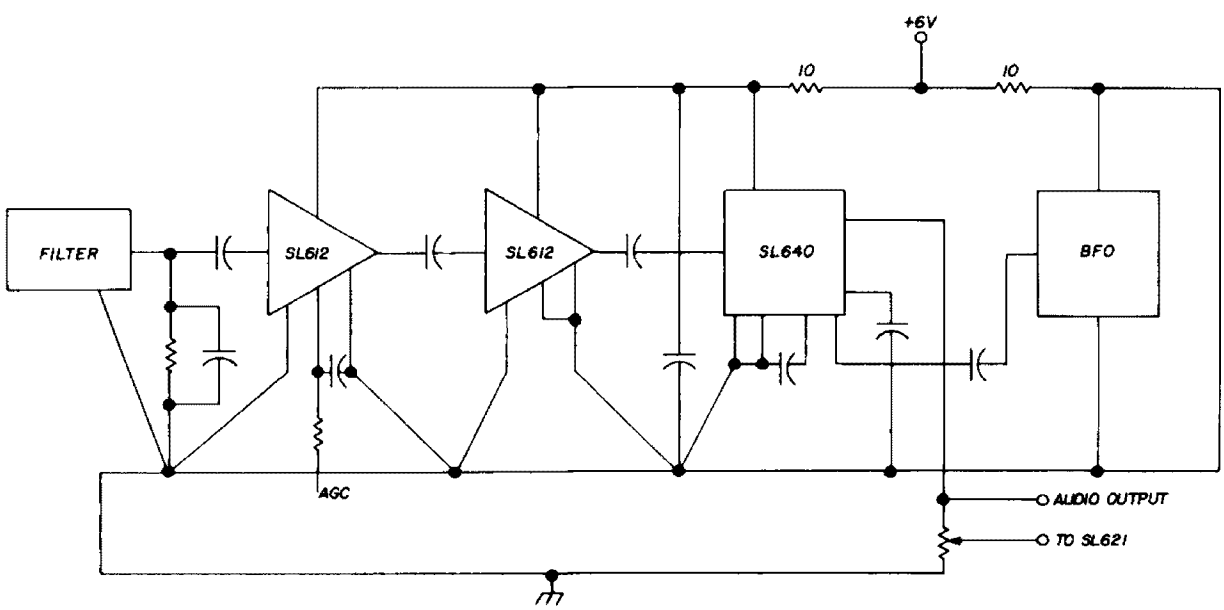


fig. 20. Proper method of grounding the ICs used in the i-f strip.

voltage-controlled gain and up to 75 mW (6-volt supply) or 200 mW (12-volt supply) output. When the SL630 is used with the SL620, which is similar to the SL621, it forms an audio agc system.[1] The SL622 is a self-contained audio agc system with an additional sidetone output which is not agc controlled. These circuits are compatible with the rest of the SL600 series, use the same power supplies and are packaged in the same T05 cases.

summary

In this article I have briefly described how the Plessey SL600-series ICs can be used in high-frequency receivers, transmitters and transceivers. It can be seen that, with the exception of power amplifiers and oscillators, high-frequency transceivers can be built with SL600 devices for all functions. Vhf and uhf transceivers can be built with SL600 ICs in all but the rf and mixer stages. These modern ICs make the design and construction of a-m and ssb equipment extremely simple, and offer the home builder the opportunity to use complex, high-performance circuits which he would not consider if he were limited to individual transistor stages.

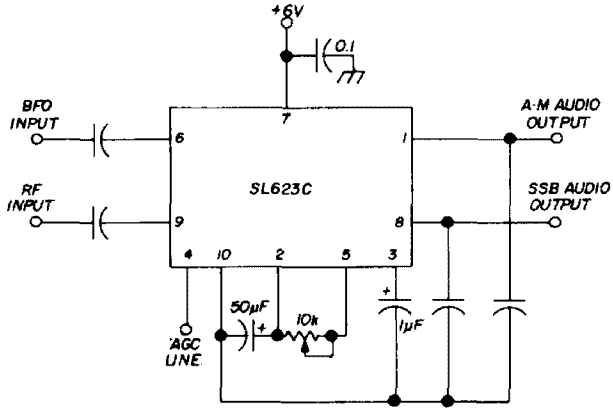


fig. 22. Suggested circuit for using the SL623C as a ssb and a-m detector.

reference

1. James Bryant, G8FNT, "Miniature Microphone Preamplifier with AGC," *ham radio,* November, 1971, page 28.

ham radio

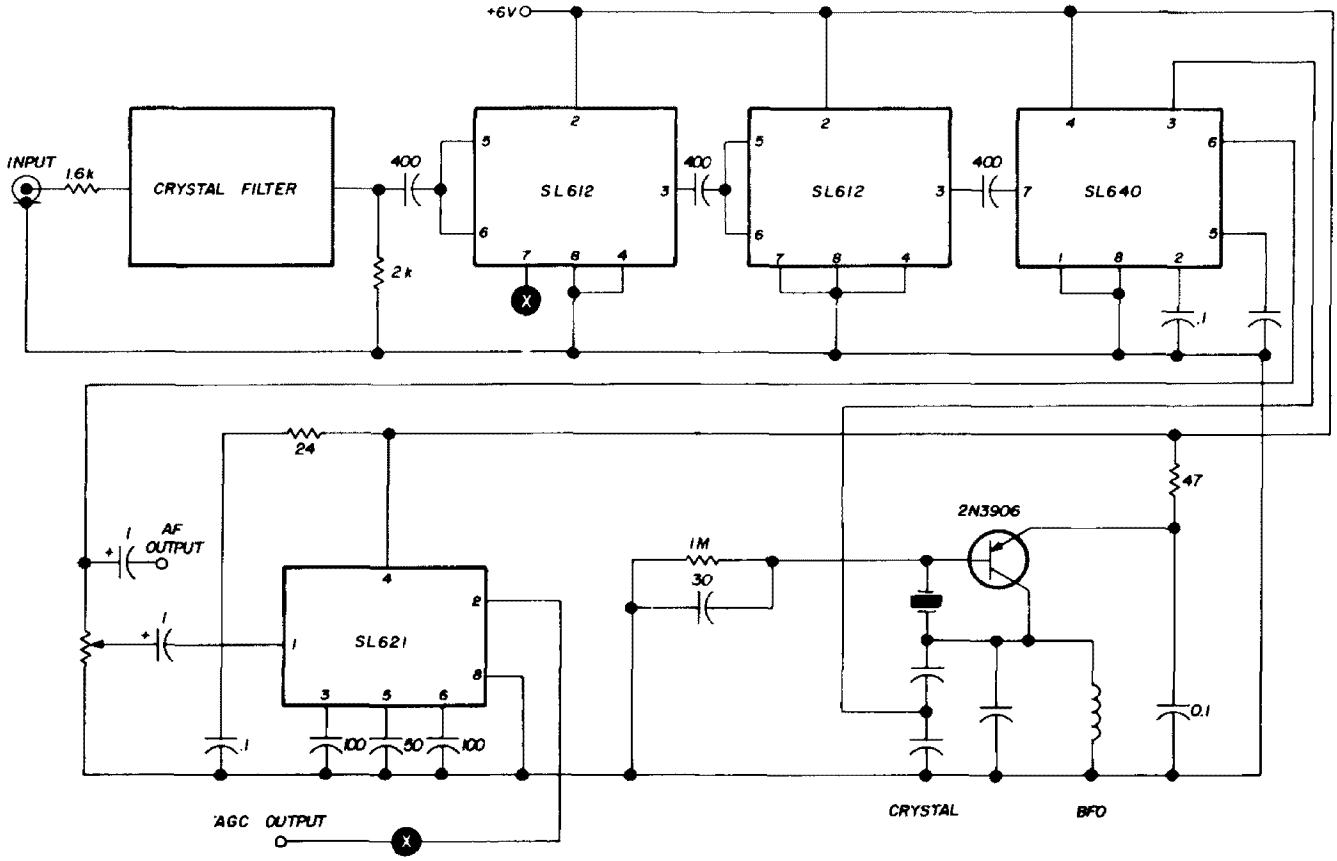


fig. 23. Complete circuit of a 10.7-MHz i-f strip, ssb detector and voice agc system.

solid-state noise blanker

Complete construction details for a solid-state noise blanker that really works

Ray Megirian, K4DHC, Box 580, Deerfield Beach, Florida 33441

In recent years there has been a rash of articles describing noise blankers and their virtues. Although the basic principles upon which the circuits depend are not new, the refinement of noise-blanker circuitry has come a long way with the help of modern solid-state devices.

Since I'm a homebrew nut of the first order, every time an article appeared dealing with the construction of a noise blanker, I'd drop everything and start breadboarding. More often than not, the end results left much to be desired. Usually, the old fashioned noise clipper did a better job with far fewer components. This frustration soon led me to the point where I was ready to climb the walls. Just to play it safe, I went back to building sweep generators, counters and receivers without noise blankers.

As this rest cure went into it's second year, I started to itch again. I found myself thumbing through my magazine collection and decided to give it another try.

the circuit

Nearly two years ago W2EGH described his version of a noise blanker, patterned after a vacuum-tube unit developed by the R.L. Drake Company.[1] This led me to assume that there was still hope since better brains than mine had done most of the hard part.

After the project finished me, I decided not to go back on the rest cure but to chance a few more weeks of cut-and-try orgies. In looking over some

of my old notes, I came across some ideas I had tried several years ago, using the RCA CA3028 IC, which had met with moderate success. Perhaps a combination of parts from each circuit could be the answer.

The result was a noise blanker that worked. At least it did for me, and I figured it might for someone else. At any rate, I submit the circuit in fig. 1 which consists of a slightly modified version of the Drake pulse-forming circuitry with an added gate transistor which shuts off the CA 3028 i-f amplifier upon receipt of a noise pulse. The blanker input is connected to the receiver mixer output and operates at 455 kHz.

The Murata dual-ceramic filter located between T1 and the input to the CA3028 is not there to provide additional selectivity, though it does that too. It is there because I accidentally discovered that this component delayed the i-f signal just about long enough to assure coincidence of the gating pulse with the noise pulse it is supposed to cancel. The home-brew receiver I used in my experiments fortunately had these filters between stages, and it took my nimble brain only a week or so to discover why the blanker worked when haywired into this i-f system but wouldn't work when mounted up with it's own i-f stage less the filter.

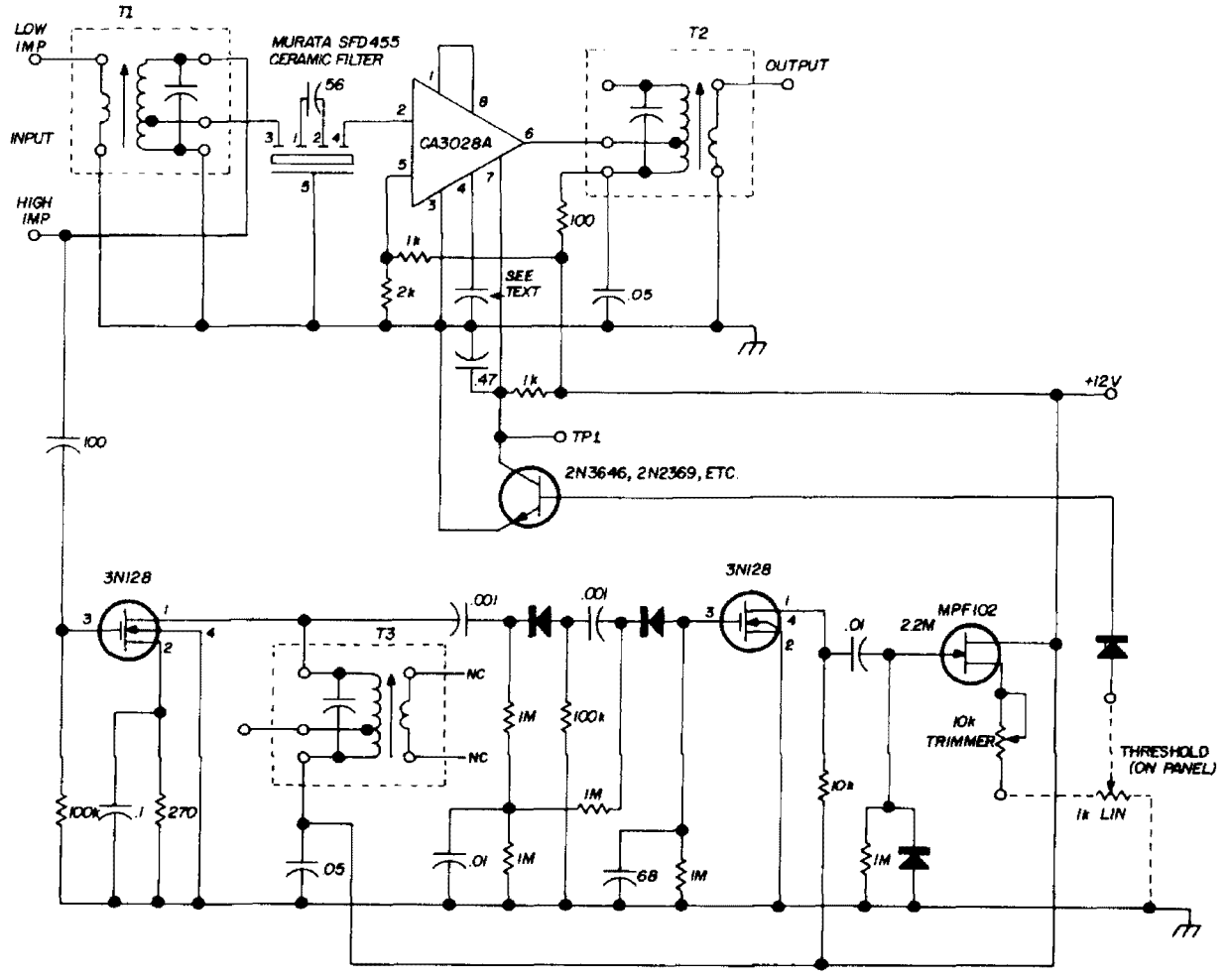


fig. 1. Circuit diagram of the solid-state noise blanker. All diodes are 1N914.

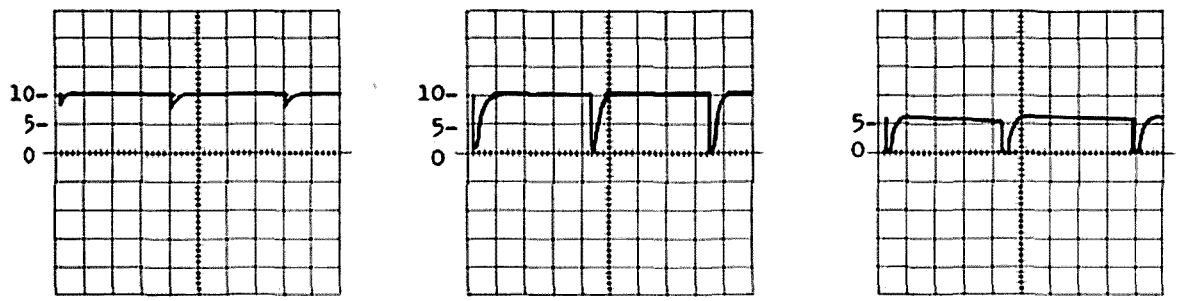
construction

A PC layout is shown for the benefit of those desiring to duplicate the circuit. The i-f transformers can be found hanging on pegs at your local Radio Shack store. I used white core transformers for T1 and T2. Either the black or yellow core units may be used for T3.

Use a switching transistor with low saturation voltage for the gate transistor. The 2N2369 and 2N3646 were both used

here and worked well. The MPF102 fet may give some trouble due to variations in characteristics, and more than one may be needed to obtain a unit suitable for use in this circuit.

threshold control at full counter-clockwise (no blanking). With a 12-volt supply the level at TP1 should be around 11 volts. Turn the threshold control fully clockwise (maximum blanking). This may



Threshold control slightly advanced.

Threshold control about one-third advanced.

Threshold control full on.

fig. 3. Typical performance of the solid-state noise blanker.

alignment

Transformers T1 and T2 are peaked for maximum signal in the normal manner. Gain of the CA3028 i-f amplifier may be adjusted to suit your particular requirements by choice of the bypass capacitor from pin 4 to ground. If you already have sufficient i-f gain in your receiver, leave it out altogether. If you need extra gain, try values between .001 and 0,1 μF.

The 10k trimmer and T3 are best adjusted while observing the gate output at TP1 on a dc scope. Start with the trimmer fully clockwise and the panel or may not cause the level to drop at TP1. The desired dc level at this time is 5 volts. Use the 10k trimmer to make this adjustment. If things don't work out as prescribed, try a different MPF102.

When I first started experimenting with blankers I built an ignition noise synthesizer, and found it invaluable for testing. However, one of the small toy battery motors at idling speed will generate some nice noise pulses for test purposes. Couple this source right to the antenna. With the threshold at minimum there should be no blanking pulses present at TP1.

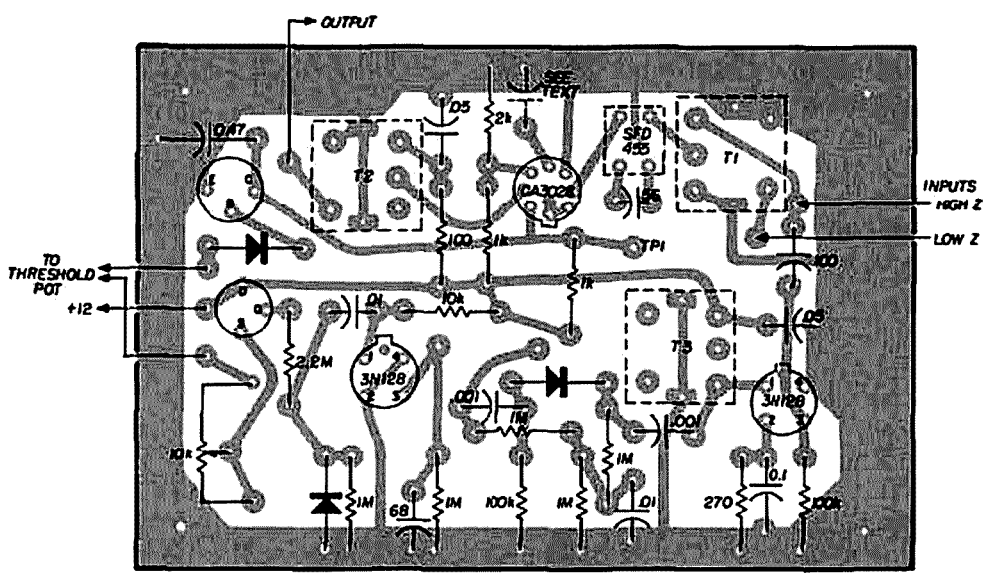
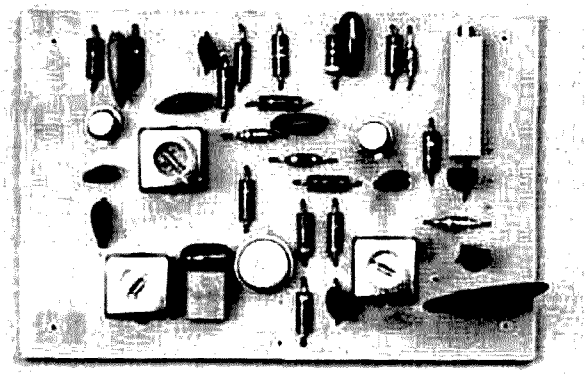


fig. 2. Full-size printed-circuit board for the noise blanker.

As the control is advenced, negative-going pulses will appear at TP1. When these pulses stretch all the way to ground, blanking of the i-f amplifier commences. Fig. 3 shows typical displays for various settings of the threshold control. Peak T3 for best sensitivity.

conclusion

Now that it's my turn to make brash statements, let me say that noise blanker performance on impulse noise is excellent, changing unreadable signals into perfect copy. Of course, you've heard all that before. The way I feel, though, is that if enough of this type of circuitry is published, perhaps eventually we'll be able to take the best from each and end



Overall photo of the assembly

up with what we've always considered to be the ultimate in performance, or at least a reasonable facsimile.

Since some of the components may be a problem with prospective builders, I will try to help out if the demand is sufficient. I refer specifically to the Murata filter, PC board and Beckman trimmer. The rest of the components are quite common and easily obtained.

I would like to hear from anyone interested in exchanging ideas along the lines presented above – if you're not on the cure.

reference

1. Frank Van Zant, W2EGH, "A Solid-State Noise Blanker," *QST*, July, 1971, page 20.

ham radio

DIODES

PIV	TOP-HAT 1.5 AMP	EPOXY 1.5 AMP	EPOXY 3 AMP	STUD-MOUNT 6 AMP
50	.04	.06	.12	.15
100	.06	.08	.16	.20
200	.08	.10	.20	.25
400	.12	.14	.28	.50
600	.14	.16	.32	.58
800		.20	.40	.65
1000		.24	.48	.75

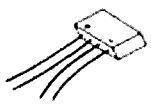
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a simple receiver-demodulator for RTTY net operation

Some ideas for a simple but effective receiver/demodulator for unattended RTTY net operation

Larry Walrod, VE7BRK, 1644 Richter Street, Kelowna, British Columbia, Canada

As more and more amateurs become interested in radio teletype and more sequential selectors become available there is bound to be an increasing interest in 24-hour standby circuits in which any amateur, so equipped, can send RTTY to any other station on the same net. This is possible even when the receiving station's operator is away from his station. To avoid tying up the regular station equipment, it is desirable to have separate gear for the net frequency. I would like to suggest an inexpensive way to set up the receiving and demodulating functions in one unit, separate from the normal station receiver and terminal unit.

You will see in fig. 1 that I have included all the *necessary* functions for reception and demodulation. However, I have eliminated a number of the circuits usually associated with each process by combining functions in a single unit.

After building forty ssb transceivers while on a recent five-year tour of the Republic of the Philippines, I got to respect the direct-conversion receiver. Although career commitments prevented my building a finished version of this unit, I feel confident that the design suggested here will be adequate for the intended purpose.

The circuit is easy to construct with only three rf coils and two mixer input

coils. All other components are subject only to low frequencies and require no special work or critical placement on the circuit board. The design, though simple, has one advantage over many more sophisticated designs as this receiver only copies the mark and space signals. Any garbage between these signals does not get through the receiver as would be the case in a more conventional audio band pass filter type design.

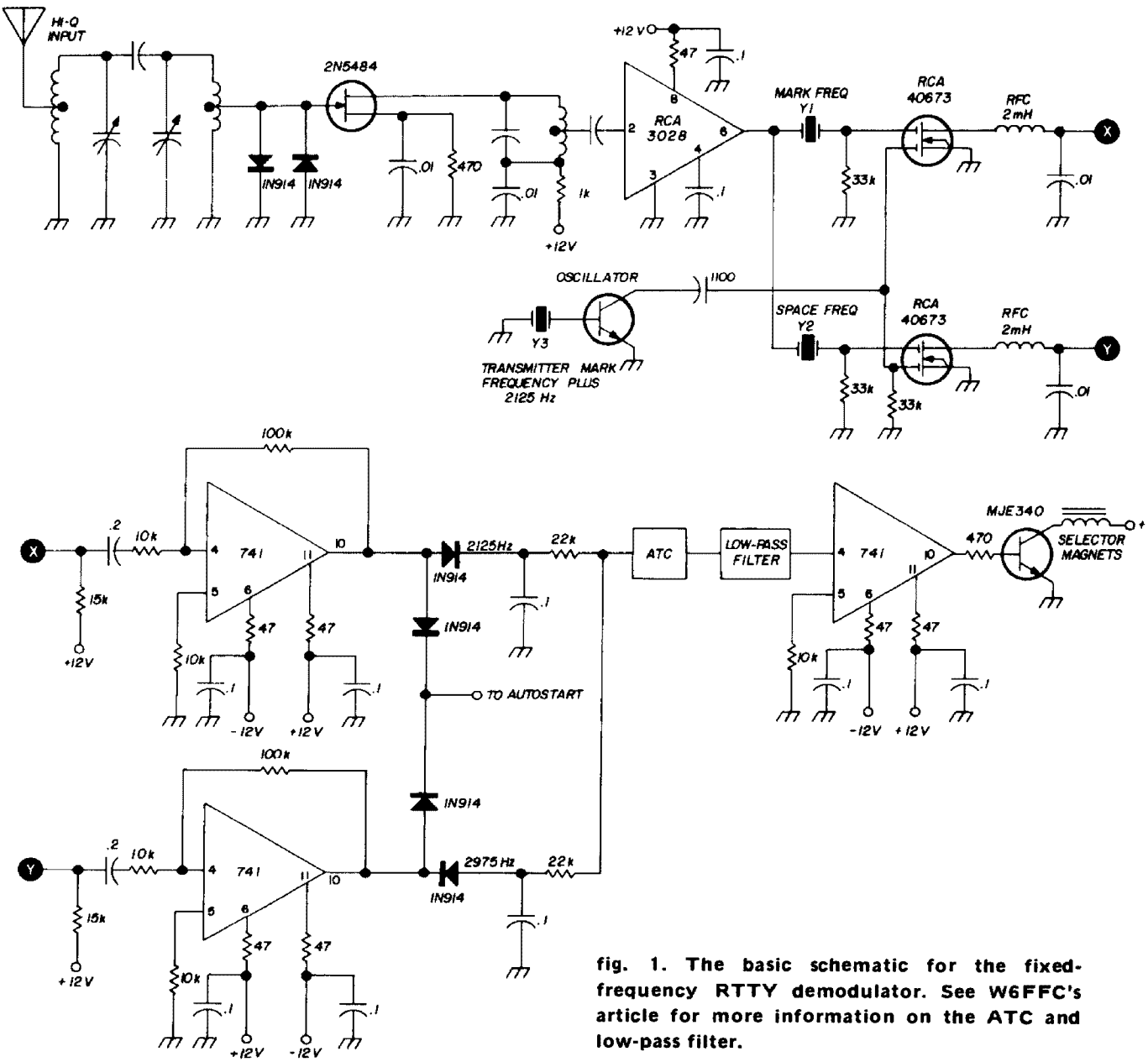


fig. 1. The basic schematic for the fixed-frequency RTTY demodulator. See W6FFC's article for more information on the ATC and low-pass filter.

crystals

It is interesting to note that the local oscillator frequency (about 2125 Hz above the transmitted mark frequency) could vary a considerable degree around the selected frequency without adversely affecting the operation of this unit. No oven would be necessary and almost any bargain priced crystal would do. This is not the case, though, with the two mark and space frequency crystals in the filters at the input to the mixers. They have to be precisely adjusted and controlled — by an oven.

The cost of obtaining accurate crystals (±20 Hz) might be a problem, but an enterprising amateur can easily get HC-6U crystals about 500 or 1000 Hz below the

required frequency and put them on frequency. Remove the metal top by mounting the two pins in a vise. While grasping the top of the case with a pair of pliers have a friend play a small propane torch flame on both sides of the case. It will be only about two seconds before the solder at the base of the case is melted enough for the top to be removed (straight up to avoid damaging the crystal).

The crystal can be inserted in an oscillating circuit and the frequency can be adjusted upwards as much as 5 kHz by very easy rubbing with a little fine sandpaper mounted over the rubber

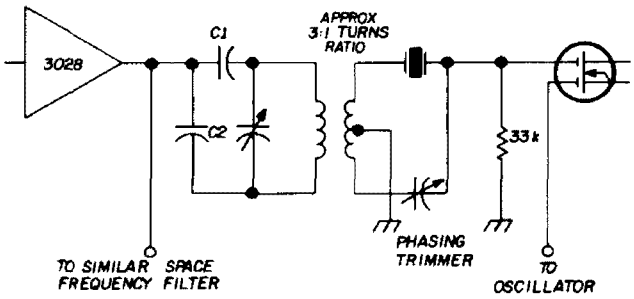


fig. 2. Suggested filter circuit for use in the circuit of fig. 1. Capacitor C1 is 15 pF and capacitor C2 is 62 pF for use on the 7-MHz band.

eraser on an ordinary pencil. Apply the sandpaper only to the plating — not on the quartz. Work a little on each side of the crystal. I have found that some crystals actually improve in activity with this treatment. Be sure to blow off any dust before making frequency checks.

Remember, though, that replacing the cover and connecting the cover to ground will have quite some effect on the final frequency. I found that grounding the case shifted the oscillating frequency of most 7-MHz crystals down about 75 Hz. If the unit is eventually to be grounded, it should be checked with the case grounded. However, when grinding and testing, you do not have to reseal the HC-6/U case each time you want to test the oscillating frequency. Just replace the cover and try it.

Also, it is not absolutely necessary to reseal the case at the end of the procedure if the crystal is going to be used in a reasonably constant atmosphere. Sealing, of course, will give far better long-term stability.

Resealing is not hard, though. There is a small vent hole near the top of the case which should be opened with a small drill and sealed *last*. Excessive heat while soldering the case to the base can lower the crystal's operating frequency by about 100 Hz. I've found a fairly effective technique is to clamp the crystal pins in a vise, install the cover (heat sinked with a large clothes pin to which a couple of brass plates have been attached) and then solder just about one quarter of the base at a time — allowing the assembly to cool between the solderings. With a little experience on this job a fellow can allow for this shift and solder the base of the case all at once (but still using the heat sink, of course). I've built about a hundred crystal filters — at first the work is very slow but it becomes easier with experience.

The balance of the circuitry is so basic and simple that it doesn't need any further treatment in this article.

reference

1. Irvin M. Hoff, W6FFC, "Mainline ST-6 RTTY Demodulator," *ham radio*, January, 1971, page 3.

ham radio



"I spent $1572.39 getting set up, and now I can't think of a thing to say."

grid-current monitor for the Heath HW-100 and HW-101

A simple circuit modification for adding a grid-current meter to your HW-100

Bob Snow, K4MFR, 713 Quarterstaff Road, Winston-Salem, North Carolina 27104

The **Heath SB-line of transceivers** has a meter-monitoring advantage over the less-expensive HW-100 and HW-101. It is obvious that several items have been deleted from the HW series in order to cut the price but it is not necessarily detrimental to the operation of the equipment to have fewer monitor points. It would probably be more correct to say that the SB-line incorporates some luxury features over the HW-line.

For example, the HW series uses a three-position slide switch for monitoring plate current, relative power output and ALC. This is compared to a five-position rotary switch on the SB-line which monitors ALC, plate current, plate voltage, relative power output and grid current. The HWs are missing the high voltage and the grid current measurement.

High voltage is of less importance since the power supply puts out all it can, and it's either there or it isn't! But the grid current is a good indication of ALC operation and a definite asset in tuning and alignment. You can have this monitoring point for just a little time, a little wire and a switch.

installation

By drilling a small hole in the front panel just below the meter and between the two slide switches, and installing a dpdt switch, you can effectively lift the meter out of its regular circuit and insert

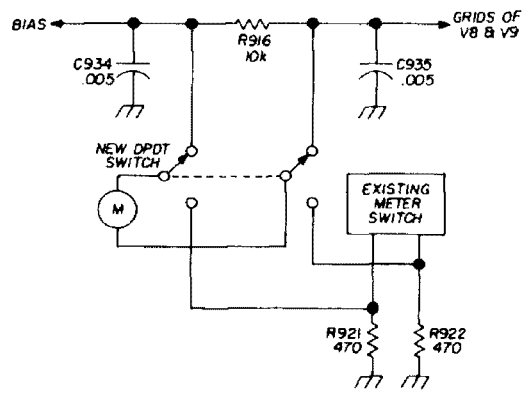


fig. 1. Grid-current monitor for the Heath HW-100 or HW-101. Only added part is S1, a dpdt switch.

it into the grid circuit as shown in **fig. 1.** Resistor R916 does not appear on a printed board. It is located on the underside of the chassis on terminal strip BR near tube V8 of the final amplifier.

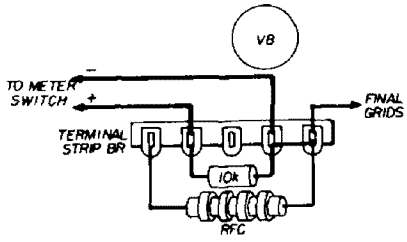


fig. 2. Location of resistor R916 under the chassis. Two new wires must be run to the added meter switch.

It is necessary to remove the cover that shields the coil assembly in order to gain access to the terminal strip. When you have located the resistor, solder a wire to each end as shown in **fig. 2.** Make sure the wires are long enough to reach to the meter. Route the pair of wires along the wiring harness, through the rubber grommet, to the other side of the under-chassis. Continue following along the harness to the front of the unit and feed the wires up through the hole in the board below the meter assembly.

Before connecting the pair of wires to the meter switch, check continuity from the resistor to code the wires for correct polarity, *see* **figs. 2** and **3.** Connect the wires as shown. Remove the connections from the meter and transfer them to the switch, observing proper polarity. Then connect the meter. It will be easier to make this installation if the meter is removed from the panel hole until all connections are made.

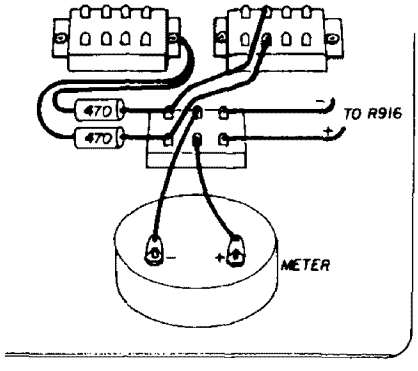


fig. 3. Modified wiring of the meter and panel switches.

The switch should be one of the miniature types, either toggle or pushbutton. The momentary pushbutton is perhaps the most desirable since the added circuitry is only effective with the pushbutton depressed. At all other times the circuit remains normal.

operation

With the new switch actuated, the meter indicates grid drive, which is very helpful during tuneup. The indication is much sharper than previously when the transmitter is in the *tune* position and adjustments are being made with the *drive preselector.* The grid meter also provides an indication of ALC action during ssb transmission. No grid current (except occasional peaks) will be indicated during ordinary transmission if the ALC is operating properly.

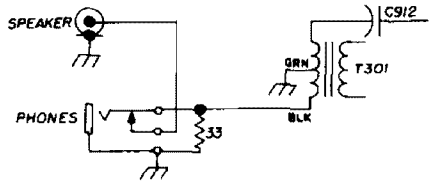


fig. 4. Simple circuit for silencing the HW-100 speaker when the headphones are plugged in.

Another annoying item in the HW-100 is the system of converting from speaker operation to headphone operation. The original circuitry connects the headphones in series with the speaker when the headphones are inserted in the phone jack. This is satisfactory as far as the headphones are concerned, but the speaker is not completely silenced. The circuit in fig. **4** shows a quick fix. It requires a 33-ohm resistor to be added at the phone jack and a length of wire to run between the phone jack and the 8-ohm speaker jack at the rear of the chassis. Rewiring the jack and grounding the output transformer as shown does the trick.

ham radio

integrated-circuit audio oscillator

This tunable wide-range solid-state audio oscillator covers the frequency range from 15 Hz to 40 kHz

Hank Olson, W6GXN, Box 339, Menlo Park, California 94025

For years I have had, sitting on the floor beside my workbench, a fine old Hewlett Packard Model 300 harmonic wave analyzer. It never was really used much because more modern wave analyzers were available at work which required less nulling and twiddling. The sheer beauty of the old HP-300 in its huge varnished oak cabinet was the only thing that prevented its being scrapped out.

Finally, in desperation for more work-shop space, the HP-300 was reduced to its component parts and carefully sorted into the various bins that make up my junk box. The cabinet was even used — as firewood (it turned out to be oak *veneer*). The largest and nicest component remaining was the dial assembly, drive mechanism and dual-variable capacitor unit. This trio had originally been the tuning of the HP-300, part of a Wien bridge oscillator covering 20 kHz to 40 kHz.

The more I looked at this marvel of mechanics, the more I conjured up pictures of an all solid-state audio oscillator built around it. One fine day, the design pieces — tuning assembly, cabinet, and integrated circuits — all fit together into a generator I have sentimentally called the HO-200. The original HO-200 is shown in fig. 1 with its old HP-300 dial (turned

down to remove the original etched calibration, and so it would fit on the front panel of the Bud C994 cabinet).

Since few amateurs are likely to have an old HP-300 tuning assembly, work was undertaken to build a second HO-200 that used readily obtainable components. The dual tuning capacitor for the second unit is made by the English firm of Jackson Brothers, Ltd., and is available here.* The vernier drive was made up from an old National *Velvet Vernier,* originally stripped from a tuning unit of a surplus transmitter. Such vernier drives are quite common and are integral parts of the National ACN and MCN dials. The basic planetary drives are also available as National parts AN-250 or AVD-250. The original dial of the vernier drive was discarded and a 5-inch aluminum disc fabricated to replace it. The finished dial is very similar to that of that used on the original HO-200, as can be seen in fig. 2.

the circuit

The HO-200 circuit (fig. 3) uses modern integrated circuits for its gain blocks. The μA740 of Fairchild is an operational amplifier in monolithic form which has fet inputs. The fet input feature allows the use of the large resistance values necessary when you design for variable-capacitor tuning of a Wien bridge.

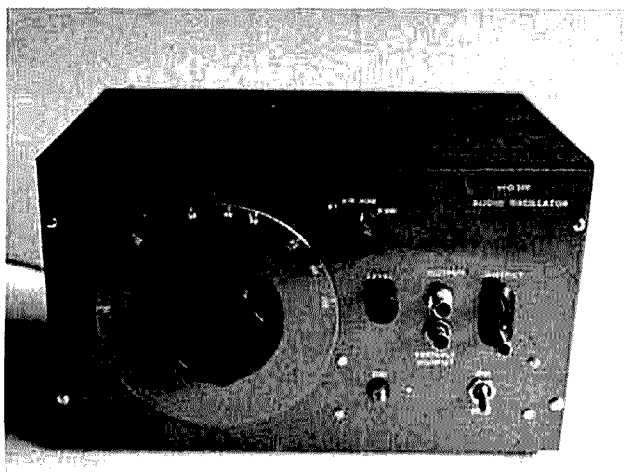


fig. 1. The original HO-200 audio oscillator used a tuning dial from an old HP-300 harmonic wave analyzer.

The typical input impedance of a μA740 is 10^{11} ohms.

A low output impedance is provided by combining the μA740 with a Motorola MC1438R op amp follower. This power IC is packaged in a small diamond case

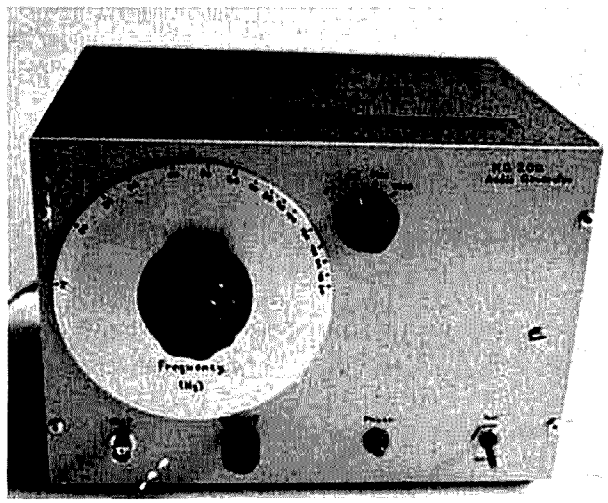


fig. 2. New HO-200 audio oscillator uses readily available parts.

for heat dissipation. The output impedance of the MC1438R is only 10 ohms, so it is capable of putting out considerable current. The MC1438R is used, as it was intended to be used, *inside* the closed loop with the op amp. (That is, the feedback to the top of the bridge is from the output of the MC1438R rather than from the output of the op amp alone.)

The power supply, which provides plus and minus 15 volts, also uses an IC – a dual regulator. The Motorola MC1468G assures that the oscillator is supplied a pair of *equal,* well-regulated voltages. The 2N5191 and 2N5194 transistors associated with the MC1468G in this dual-regulator circuit pass the major part of the current, as is done in most IC regulator circuits.

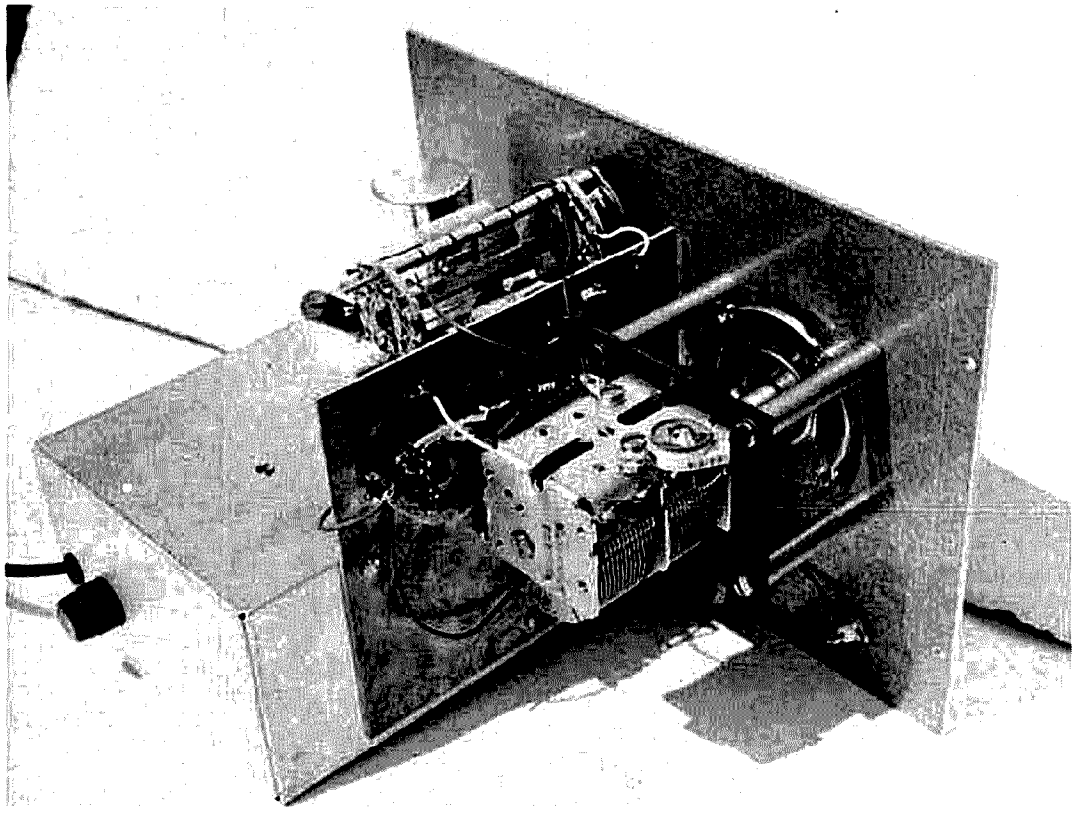
The non-linear resistance which is used to stabilize the amplitude of oscillation is not a lamp bulb, as is most common in

*The Jackson Brothers 5084/2/518H dual 518-pF tuning capacitor is available from M. Swedgal, 258 Broadway, New York, New York 10007. The price is $5.00.

Wien bridge oscillators. Instead, a thermistor is used. Of course, since a thermistor *decreases* resistance as it is heated, it must be placed in the top leg of the bridge. This is just the opposite of the lamp bulb, which *increases* resistance with temperature and must be placed in the bottom leg of the bridge. This difference is illustrated in fig. 4. Other such uses of a thermistor in a Wien bridge is described in references 1, 2 and 3.

Centralab PA-300 and two PA-0 sections. The two switch wafers were spaced about 2-3/4 inches apart to allow room for the resistors. These ceramic wafer-switch sections appear to be of sufficiently good insulation for use with the high values of resistance involved. If you use series resistors as I did, teflon or ceramic standoff insulators could be used for additional terminals.

Since the common terminal of the



Inside view of the audio oscillator showing the tuning capacitor, range switch and copper-clad board used as a ground plane.

construction

The ranges that the HO-200 will cover are 15 to 200 Hz, 150 Hz to 2 kHz, 1.5 to 20 kHz, and 3 to 40 kHz. Range switching is accomplished by changing resistance values in the positive feedback side of the bridge. It is important to use 1% resistors here; I used deposited-carbon types. I used some series pairs of resistors to make up resistance values in order to avoid buying special values.

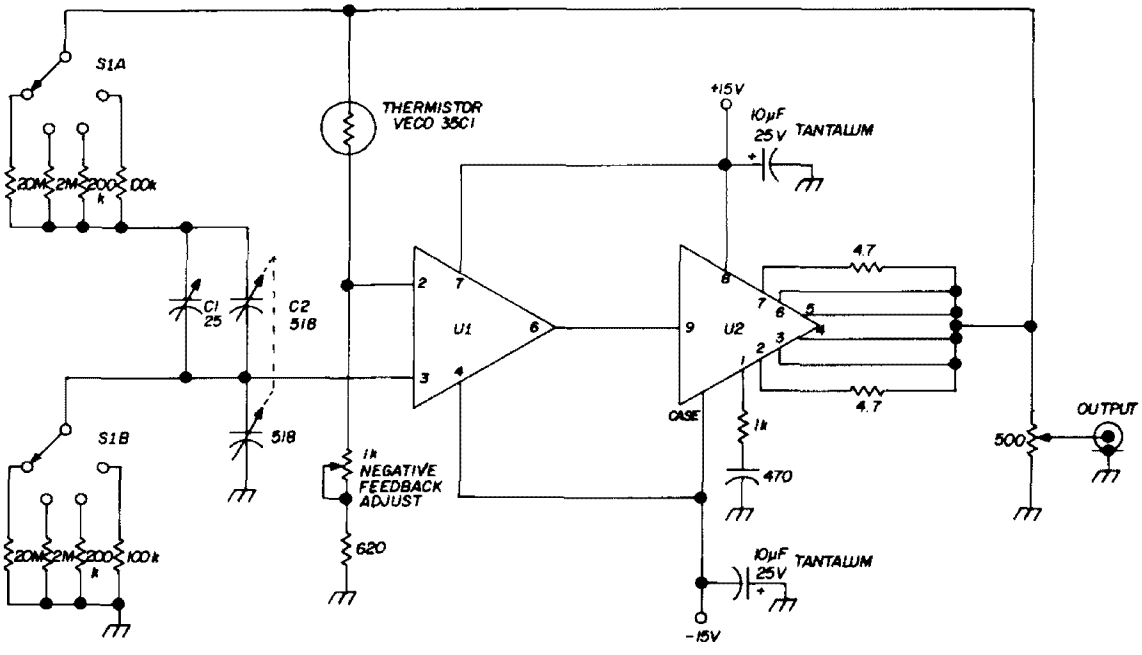
The resistors were mounted by soldering them to the wafer terminals of a two-wafer switch assembly made up of a

dual tuning capacitor is *not* grounded, it must be mounted in such a way that it floats above ground. I accomplished this by mounting the tuning capacitor body on a square of phenolic, and using an insulated shaft coupling for tuning. The phenolic was mechanically supported by four two-inch spacers screwed to the front panel of the HO-200.

The actual oscillator circuitry was built on a piece of copper-clad laminate, such as is used to fabricate etched-circuit boards. In this case, double-sided copper-clad board was used. The part of the

board that is above and next to the chassis is of special interest. One side of the laminate is at -15 volts and serves as a heat sink for U2. This "hot" portion of one side of the double-sided copper-clad laminate is isolated from the lower portion (which is bolted to the chassis) by a strip of insulation where the copper foil was peeled off. This is easily done with a rule and an X-acto knife or razor blade.

The other side of the board is entirely at ground potential and is the side toward the tuning capacitor where the bulk of circuitry is mounted. Several small insulated standoffs are used to support components on the grounded side of the board; they are fixed to the board by wetting their metal bases with solder and soldering them to the copper-foil ground plane. As can be seen from the photo of the inside of the HO-200 there are really very few components to mount; there-



C1 4.5 - 25 pF trimmer capacitor

C2 Dual 518-pF tuning capacitor (Jackson Brothers 5084/2/518HO)

U1 Fairchild μA740C, Signetics NE536T, National NH0042C or Intersil ICL8007C

U2 Motorola MC1438R or MC1538R

fig. 3. Circuit of the solid-state Wienbridge audio oscillator uses two ICs. U1 is a high input impedance operational amplifier; U2 is a low-impedance output op amp follower.

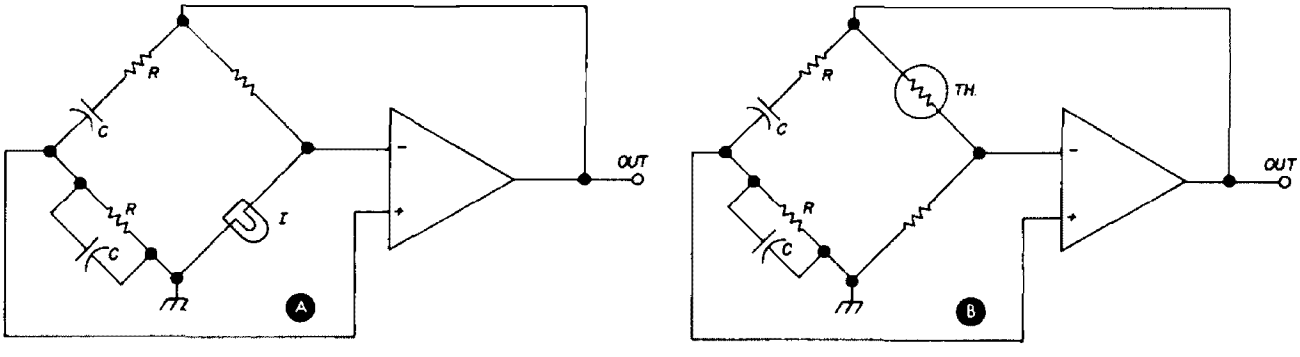


fig. 4. Use of nonlinear feedback elements in the Wien bridge circuit. The circuit in (A) uses the traditional incandescent lamp. The circuit in (B) uses a thermistor. See text.

fore, an etched circuit board was not deemed necessary.

There is one more consideration that must not be overlooked. Because the common terminal (body) of the dual tuning capacitor floats above ground, it has some finite capacitance to ground — measured to be about 10 pF in my unit. Since this is effectively in shunt with the parallel R-C branch of the bridge, a similar capacitance must be added across the tuning capacitor in the series branch adjusted by means of the trimpot in the negative feed-back side of the bridge.

The regulated power supply is entirely located on the underside of the 5x7 chassis. This is to confine 60 Hz *electric* fields so that they will not be picked up by the high impedance circuits in the Wien bridge. One electrolytic, the positive supply filter capacitor, protrudes through the top of the chassis; but its can is grounded and so has no external 60 Hz field on it. An *aluminum* 5x7 chassis is

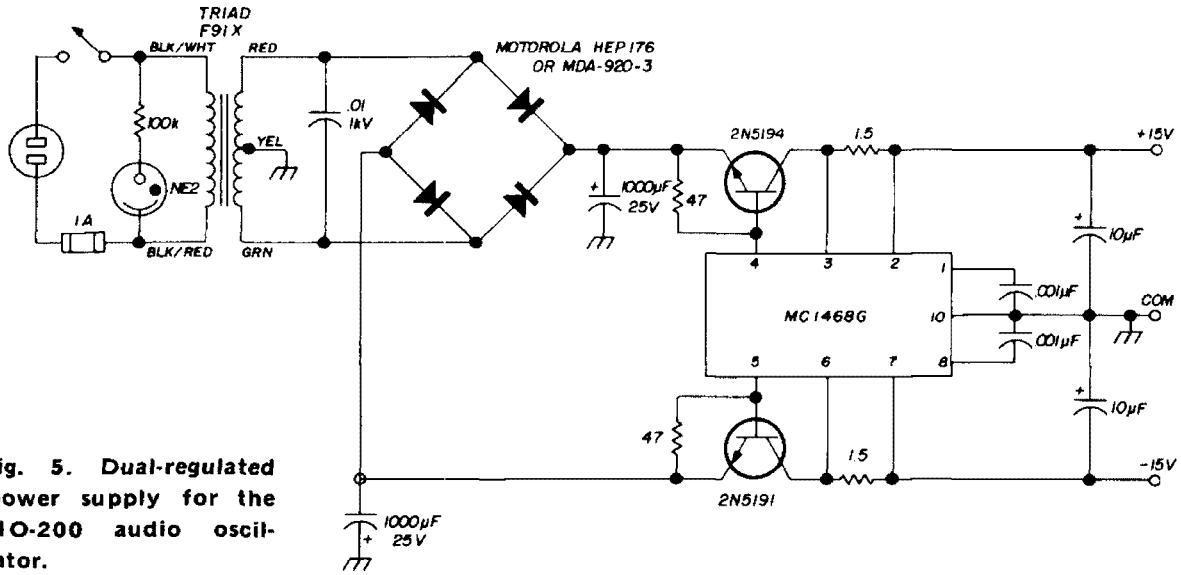


fig. 5. Dual-regulated power supply for the HO-200 audio oscillator.

of the bridge. C1 is a variable trimmer for this purpose, and should be set at about half capacitance for a start.

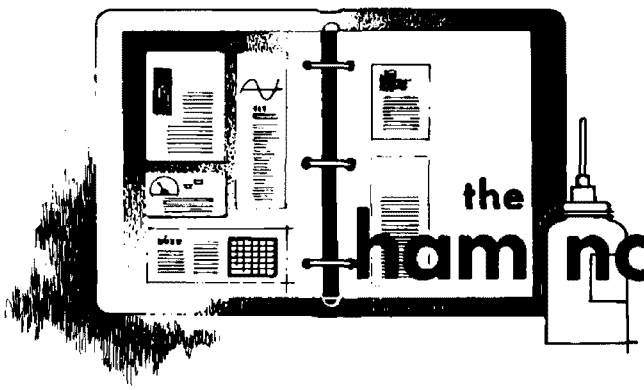
With the tuning dial at lowest frequency (x10 scale), measure the output voltage. Then tune to maximum frequency (x10 scale) and adjust C1 to give the same output voltage reading. The idea here is that 10 pF stray capacitance will not be much (2%) unbalance out of 518 pF at the low-frequency end (maximum capacitance), but will upset the matching of capacitance when the tuning capacitor sections are at their minimum (maximum frequency end). Therefore, C1 is best adjusted at the high-frequency end.

There is one other adjustment that can be made to the HO-200. With the output pot at maximum, you should see about 16 volts p-p at the output jack. If this amplitude is some other value, it may be adequate here since we are only concerned with *electric* fields (not magnetic fields, too). If you want to see the great sensitivity of the circuit to stray 60-Hz fields, try operating the HO-200 out of its cabinet on the X1 scale. You will note (especially near 60 Hz and 120 Hz) severe beat phenomenon and a generally hopeless mess of waveforms.

references

1. R. Fulks, "High Performance, Low Cost Audio Oscillator with Solid-State Circuitry," *GR Experimenter,* August-September, 1962, page 15.
2. I.C. Zero, "Solid-State Wien Bridge Audio Oscillator," *Audio,* July, 1965, page 19.
3. H. Olson, "Dual Op Amp Makes Simple Sine-Wave Generator," The Electronic Engineer, August, 1971, page 70.

ham radio



logic test probe

As a computer engineer in a small company, I sometimes have to build my own test equipment. One of the most useful pieces I have come up with for trouble-shooting digital circuits is a test probe with memory that responds to either positive or negative going pulses and/or levels. Since I work only with TTL logic, the probe is designed to work with this. Because of the memory function, it spots the presence of single control pulses with just one operation of the circuit

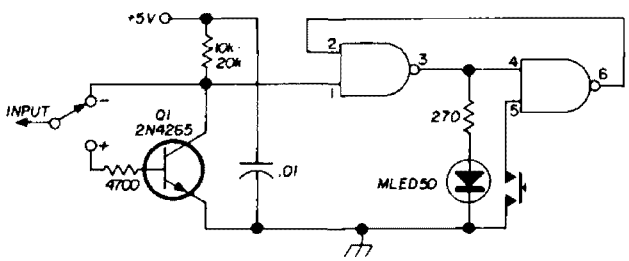


fig. 1. Circuit diagram of the logic test probe. Use IC logic of the same type you are testing, RTL, DTL, TTL, etc.

under test, while with a scope I have to trigger the circuit many times to clearly determine the presence of the control pulse. In addition, a high or low state is easily seen, simply by touching the desired point. It may be hard to believe, but I can go all day without ever turning the scope on.

There are two switches a Memory Disable switch and a Pulse Polarity switch. Memory Disable is a push-button that resets the memory to the low state when depressed. Pulse Polarity is a toggle switch located out of the way and selects whether the probe responds to a high-level or pulse (+5v) or a low-level or pulse (ground).

The probe circuit, fig. 1, is simplicity itself, one TTL quad NAND integrated circuit, three resistors, one NPN Transistor, an LED and two switches. I used a 7400 integrated circuit, a motorola MILED 50 LED and a 2N4265 transistor. All parts are miniature and of high quality for reliable operation. Too much money has been spent for inexpensive parts to build test equipment that the owner ends up not trusting. Construction depends to a large degree on what you have on hand. I used a part from an old oscilloscope probe, a piece of 3/4-inch plastic tubing, and the plastic barrel from a ball-point pen with a piece of threaded rod melted into it. Conscientious workmanship is in order if you do as I did, and glue the two major parts together.

To use, connect the two power leads, and with the polarity switch set for positive pulses, touch the tip to the positive supply. The light should go on. Touching the tip to ground should not affect the light and only depressing the memory disable switch will let the light go off. With the polarity set to negative, touching the probe to the power supply should not cause an indication. The only way to get the light to light is to momentarily ground the probe tip. The only way to get the light off is to disable the memory.

In use, the memory switch can be held depressed and the probe will respond

only to the level touched. If an oscillator output is probed the light will dim when memory is disabled, and brighten to full brightness when the memory disable is released. This project has proven to be a real time saver and well worth the few

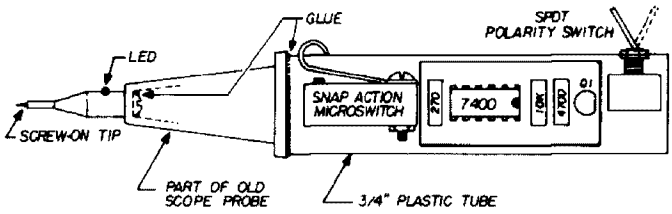


fig. 2. Construction of the logic test probe.

hours put into building it. While not everyone is deep into TTL, the applications are increasing, as I see digital repeater controls, digital circuits in RTTY applications, digital clocks, digital counters, digital volt meters and many others. This logic probe just might be the vtvm of the digital world.

Bill Rossman

frequency pre-scaler

There are a couple of things which the article on frequency scaling by F.E. Emerson in the September, 1972, issue didn't point out which may be of interest to some readers. I have been using the Fairchild 95H90 IC in a remarkably similar prescaler for some time with excellent results. My particular circuit seems to yield a little more sensitivity than Mr. Emerson observes, but perhaps I just happen to have a hotter 2N5179 in the front end.

Mr. Emerson may not have considered the consequence of using this scaler with counters that already have a good high-frequency capability such as the Heathkit IB-1101. Because the prescaler output is a square wave with quite good rise time, a counter which can count to 60 MHz or more can erroneously count the second or third harmonic of the scaled frequency. The count is reliable – it's simply wrong and is not the fault of the scaler. The simple solution to this problem is to roll off the frequency response of the scaler output. In my own prescaler I use a simple L-section filter to roll off frequencies above 30 MHz. This allows the scaler to be used up to 250 MHz and eliminates the possibility of false counting at the scaled frequency. A schematic of my own scaler, for those who are interested, is shown in fig. 3.

Ernie Guerri, W6MGI

telescoping tv masts

The majority of tv-type antenna masts currently on the market use a cotter pin under the bottom of each section that the upper section rests on. The cotter pin fits into a small slot in the upper section. However, there's always the possibility that the upper section hasn't seated properly during installation; if this happens the antenna will "windmill" back and forth, eventually wearing through the cotter pin.

As a safety measure a small hole should be drilled through *both* mast sections and an additional cotter pin installed. This prevents any rotation and could preclude renting a 40-foot ladder later to correct the problem.

Richard Mollentine, WAØKKC

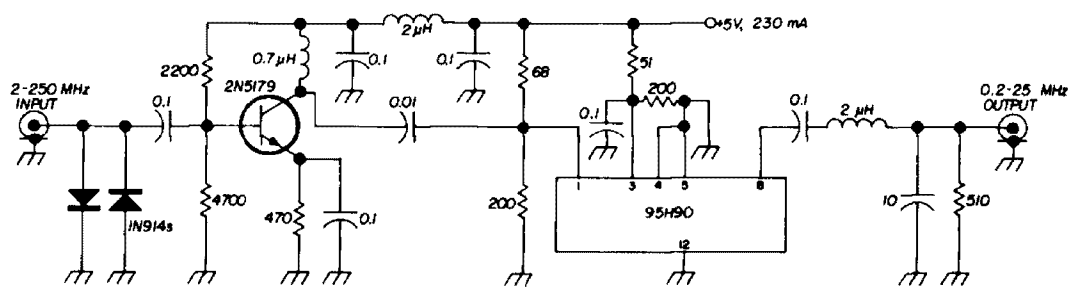
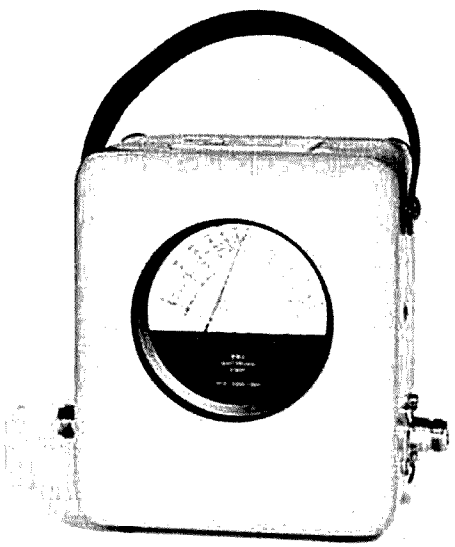


fig. 3. Frequency scaler has sensitivity of 20 mV at 175 MHz, 40 mV at 220 MHz and 90 mV at 250 MHz. All capacitors are CK06 with shortest possible leads. All unmarked pins on the 95H90 IC have no connections (do not use as tie points).

new products

rf power and vswr meter



The new Thruline® model 4342 Dual Wattmeter-VSWR Monitor displays the three prime rf transmission measurements at once on a single meter face: Forward and reflected power are indicated by individual pointers and vswr is monitored on a third scale from the intersection of the two power pointers.

Unlike most VSWR meters, the model 4342 does not require *any* adjustments to full-scale deflection, or any switching before vswr readings can be taken: The entire set of three transmission parameters is read out simultaneously during normal rf operations.

Power and frequency range of the new Bird Dual Wattmeter-VSWR Monitor depend on two plug-in elements selected from more than eighty choices available with the company's popular Thruline Model 43: Full scale power levels at ±5% accuracy range from 10 to 5000 watts for forward indication and 1 to 500 watts for reflected, in discrete frequency bands from 2 to 2300 MHz (for increased resolution, the reverse power element is ten times more sensitive than the forward power element). Model 4342 Insertion vswr in 50-ohm systems is a low 1.035 to the point of measurement. A choice of QC Quick-Change connectors permits mating with N. BNC, TNC, UHF, C, SC, LC, HN, LT, GR type 874 and 7/8" EIA lines without the need for performance-degrading adapters.

The new Bird Dual Wattmeter-VSWR Monitor is a portable instrument weighing only 5½ lbs. Model 4342 price (with female N connectors) is $265. Plug-in Elements are $30 to $75 (two required).

For more information, write to Bird Electronic Corporation, 30303 Aurora Road, Cleveland, Ohio 44139, or use *check-off* on page 94.

printed-circuit service

A valuable new service is now being offered to amateurs who build their own equipment. Most construction articles in the amateur magazines call for a circuit board. In some cases the layout is provided, but few amateurs are equipped to fabricate their own boards conveniently at home. For those who would rather concentrate on the electronics than on circuit boards, Mac McClaren, W8URX, is offering printed circuit boards, drilled and ready for parts mounting.

If he already has the artwork for the board you want, he can deliver a finished board in ten days. He offers two-week service on new items (from current magazines). He is also able to make boards

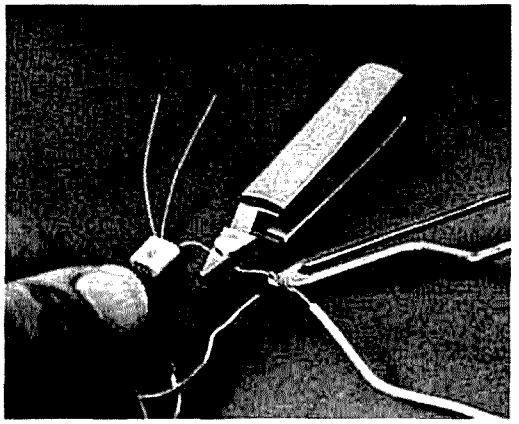
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from your artwork in three weeks -- write for a quote and full instructions.

All work is offered on a money-back guarantee. All shipments fob Cleveland. All boards use the very best material, G-10 epoxy, 1/16" thick with 2-ounce copper; all holes are drilled with solid carbide bits to assure burr-free holes.

For more information on this unique printed-circuit service, write to D.L. McClaren, W8URX, 19721 Maplewood Avenue, Cleveland, Ohio 44135, or use *check-off* on page 94.

heat sink



Xcelite has just introduced the no. 80 heat sink for absorption and dissipation of heat in soldering operations where adjacent, delicate electronic parts might be damaged by overheating. The manufacturer states that for optimum heat absorption and dissipation the jaws of the no. 80 heat sink are made of copper, with nickel plated surfaces to prevent solder adhesion. The spring-loaded gripping surfaces will not slip, yet have a smooth finish to prevent scratching fine wires. Exceptionally compact, this heat sink is 3-1/4-inches long in length and weighs only 1/2 ounce. An insulating cushion grip permits burn-free handling.

The no. 80 heat sink, made in the United States to highest quality standards, is now available through Xcelite's nation-wide local distributors at a list price of $2.40. Product Bulletin 5721 contains additional information, and may be obtained by writing to Xcelite Incorporated, Orchard Park, New York 14127, or by using *check-off* on page 94.

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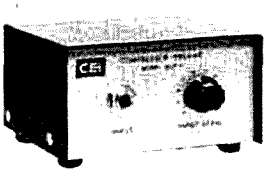
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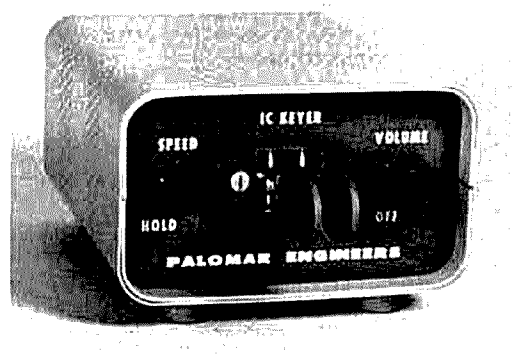
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ic keyer



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Low impedance ICs are used throughout to minimize rf pickup and they are heavily bypassed to make the keyer practically rf proof. The exclusive blocking-oscillator pulse generator gives instant start and never produces an unwanted dot or dash.

Up front is the precision Brown key mechanism that is adjustable for dot and dash travel and tension. On the rear deck is an input for receiver audio so that both the received signal and the keyer monitor tone can be heard on the built-in speaker or on headphones.

The unit sells for $87.50 postpaid in USA. For more information, write Palomar Engineers, Box 455, Escondido, California 92025, or use *check-off* on page 94.

tool kit brochure

A 16-side brochure describing field engineer tool kits and cases has been published by Jensen Tools and Alloys. The brochure describes six professional tool kits, each engineer-designed to do a particular job. Included in the publication are the JTK-17 Field Engineer Kit, which contains over 100 tools in an executive attache case; the JTK-2 Electronic Technician Kit for industrial personnel; the JTK-16 Compact "Detective" Kit, which

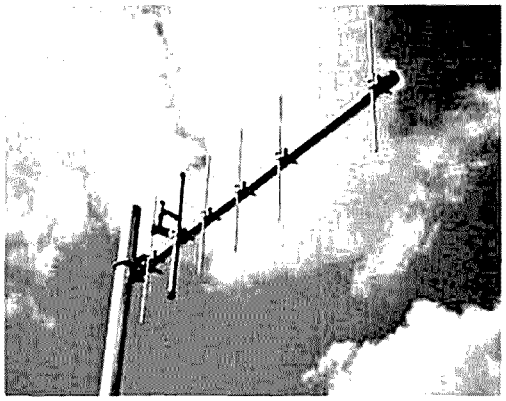
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contains 30 multi-purpose tools in a zipper case; the JTK-27/37 Electronic Lab Kit, a complete portable electronics tool kit with test instruments in two attache cases; the JTK-90 Instrument Repair Kit, emphasizing watchmakers' tools; and the JTK-80 Electronic Technician Roll-Pouch Kit, a lower priced kit for technicians, students and kit builders. Kits range in price from $44.50 for the JTK-80, to $555.00 for the JTK-27/37.

A variety of cases sold separately are described and range from a canvas-like roll-pouch to the Jensen Extra-Deep Deluxe Case which features a full six-inch depth and two removable tool pallets. A total of nine cases are presented.

To receive a copy of this brochure, contact Jensen Tools and Alloys, 4117 N. 44th Street, Phoenix, Arizona 85018, or use *check-off* on page 94.

450-MHz Yagi



Cush Craft has just introduced a new six-element, rear-mount 450-MHz Yagi for amateur fm repeater operation. It can be used for control links and stations for monitor applications and access to 450-MHz repeaters. It is priced at $10.95 amateur net and exhibits 10-dB gain. It has direct 50-ohm Reddi Match feed with built-in coax fitting. The boom is 35" long and overall weight is 3 lbs. Model No. A449-6 is available through all Cush Craft distributors. The antenna is also available for commercial service.

For more information, write to Cush Craft, 621 Hayward Street, Manchester, New Hampshire 03103, or use *check-off* on page 94.

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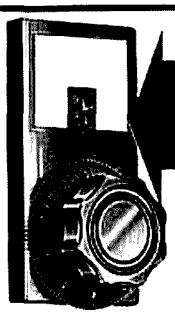
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DIGIPET 60/160

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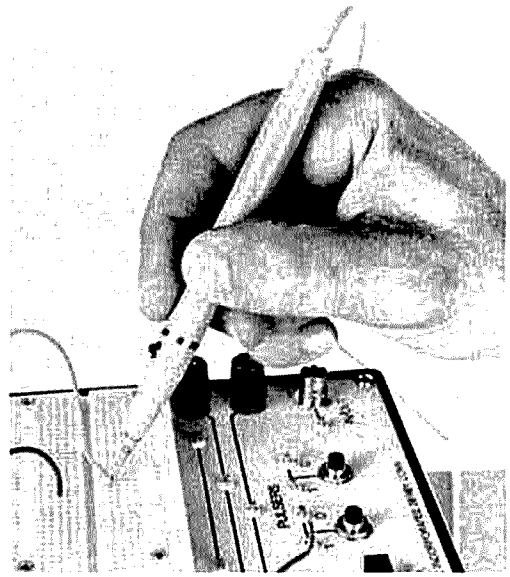
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logic-tester probe



EL Instruments announced the addition to their product line of the LT-1, Logic Tester Probe. The LT-1 is a rugged, compact pencil type instrument used to determine logic levels. It operates from the dc supply of the system under test.

The logic levels are detected by an infinite life LED used as an indicator light. The LED is on at the (+) logic 1 level and off at the 0 level. The LT-1 is a high-impedance device, therefore there is no circuit loading. The frequency response is dc to 12 MHz.

The LT-1, complete with power leads, is available from stock for $11.95. Complete information and quantity prices are available from EL Instruments, Inc., 61 First Street, Derby, Connecticut 06418 or by using *check-off* on page 94.

selective call systems

Bramco Controls Division of Lederex, Inc., offers an extensive line of selective calling encoders and decoders for attachment to mobile and base station fm equipment. The line includes units compatible with GE's Channel Guard, Motorola's Private Line and RCA's Quiet Channel selective-call systems. There are also many three, five and seven digit units compatible with the Touch-Tone systems becoming more popular daily on amateur repeaters.

More Details? CHECK–OFF Page 94

The line includes a selection of mobile decoders which will activate headlights, the car horn or a latched-on signal lamp to indicate a received call while the driver is not in the vehicle. At the top of the line is a sophisticated mobile identifier system which informs the dispatcher – by means of lights and a digital readout – of the identity of the calling mobile and if the communication is routine or an emergency.

For more information on this complete line of selective calling systems write to Bramco Controls Division, Ledex, Inc., College and South Streets, Piqua, Ohio 45356 or use *check-off* on page 94.

bird wattmeter catalog

Available now is an eight-page supplement to the 1971 general catalog which lists over thirty new Thruline® rf directional wattmeters, Termaline® rf load resistors, and a 100 watt attenuator added this year by Bird Electronic Corporation, manufacturer of instruments for rf power measurement.

Shown for the first time are 75-ohm wattmeters for uhf TV and 75-ohm loads, in addition to the usual 50-ohm equipment. The brochure introduces two new product lines: A 51,000 BTU/hour heat exchanger and an instant-output wide band 150 watt rf power source. Recognizing the substantial number of radio amateurs and boating enthusiasts in the communications industry, the supplement also shows the new economical Ham-Mate® and Marine-Mate® wattmeters.

Prices are included with equipment photos and performance specifications. General Catalog Supplement GCS-72 is available free from Bird Electronic Corporation, 30303 Aurora Road, Cleveland (Solon), Ohio 44139 or by using *check-off* on page 94.

Thruline, Termaline, Ham-Mate and Marine-Mate are all registered tradenames of the Bird Electronic Corporation.

KW ELECTRONICS LIMITED

The KW2000B the transceiver with 160 Meters

$699 with spkr and AC pwr supply



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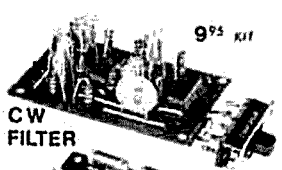


COMMAND PRODUCTIONS RADIO ENGINEERING DIVISION P.O. BOX 26348-K SAN FRANCISCO, CALIF. 94126

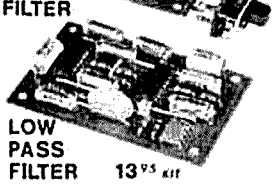
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MINIATURE SUB-AUDIBLE TONE ENCODER $14.95 Wired-Tested



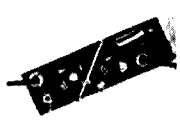
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HF 450 WIRED **29.95**



15DB GAIN WITH 4.5DB NOISE FIG. SILVER PLATED G11 APOXY GLASS P.C. BOARD HAS LOW RF LEAKAGE OPERATING RANGE 406 TO 470 MHz SIZE 3 5/8 BY 1 1/8 INCHES WILL OPERATE FROM 10 TO 15 VDC

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SPECIAL HR-2 or HR-212 WITH SCAN 2 and HF-144U PRE-AMP HR-2A $229.95 HR-212 $259.95

CRYSTALS FOR HAM USE

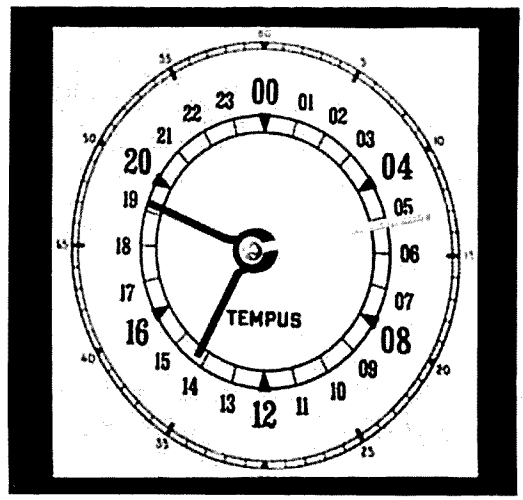
HR-2 & 212	RECEIVE	.....	$3.99
	TRANSMIT	.....	$5.75
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ALL PRICES INCLUDE SHIPPING



Topeka FM Communications Land Electronics
1313 E. 18 Terr.
Topeka, Kansas 66607

two time-zone 24-hour clocks



Tempus Instrument Corporation has introduced a 24-hour clock, powered by a high-quality four-jewel battery-operated clock movement made in West Germany. One regular C-size flashlight battery runs the clock for at least 18 months. The case is of scratch-resistant Acrylic material measuring 6 x 6 x ½ inches.

Three multi-colored hands permit simultaneous reading of hours and minutes 24 hours a day in *two* different time zones of the world. This feature makes the clock especially useful for amateur radio operators. Each clock comes with a hanger and with an aluminum bracket for use as a wall clock or desk clock.

Tempus also offers a similar model as a plain 24-hour clock with sweep second hand. All products are fully guaranteed for one year and sell for $29.50. Additional information is available from Tempus Instrument Corporation, Villa Italia Center, Suite 594, 7200 West Alameda Avenue, Denver, Colorado 80226 or by using *check-off* on page 94.

radio shack catalog

Radio Shack's new 1973 electronics catalog features the company's complete line of experimenter's components, Realistic home entertainment products, audio equipment and Citizen Band equipment, Micronta test instruments, Archer antennas, and Archerkit and Science Fair electronic and hobby kits.

More Details? CHECK—OFF Page 94

The 180-page catalog, with more than half of its pages in full color, lists thousands of electronic items such as tubes, transistors, cables, tools, connectors, wire, plugs and adapters for amateurs, hobbyists, experimenters, technicians or anyone wanting a wide selection of parts, accessories and maintenance items.

Among the products being introduced as new for 1973 are stereo and four-channel amplifiers, receivers, adapters and tape decks; speaker systems, CB radios, Archerkits, Science Fair Lab kits, stereo radios, scanning monitor receivers and a miniature electronic calculator.

Catalog 227 is available free on request from any of Radio Shack's more than 1500 stores in all 50 states and Canada, or by mail from Radio Shack, Department R-26, 2617 West Seventh Street, Fort Worth, Texas 76107 or by using *check-off* on page 94.

crt terminals

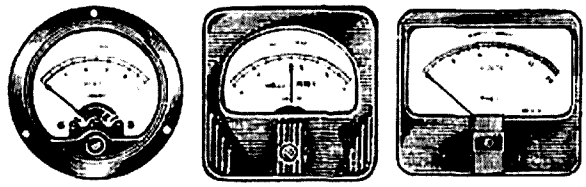
A new data sheet describing the Series 200 Video Terminal Controllers, free-standing keyboards and monitors is available from Ann Arbor Terminals, Inc. The new literature emphasizes the application of Series 200 crt units in conjunction with, or as substitutes for hardcopy printout terminals.

Features of the Series 200 equipment include noiseless operation, preview and editing capability, very high speed and exceptional reliability – features not available on electromechanical data capture terminals. Other features of Ann Arbor's line are compatibility with existing keyboards and data capture terminals and the capability for driving multiple standard TV sets or 525-line video monitors. There are many units in the line ranging from $795 to $1195 in single quantities.

For the new data sheet write to Ann Arbor Terminals, Inc., 6107 Jackson Road, Ann Arbor, Michigan 48103 or use *check-off* on page 94.

FROM YOUR COST CUTTER...

METERS!



5000 IN STOCK

FMers: 0-Center, Micro Amps, Various Scales — 2½" — $3.95; 3½" — $4.95

Also tubes, transformers, semi-conductors relays, capacitors and test equipment.

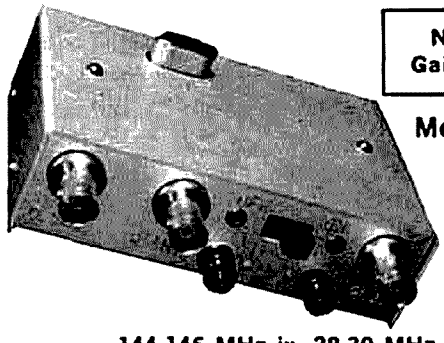
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. . . THE BEST 2 METER CONVERTER



Now with Gain Control

Model 407-A

$49.95

ppd.

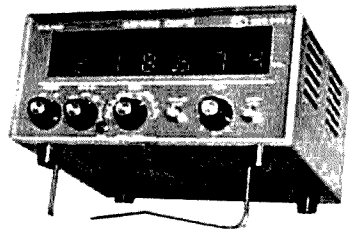
144-146 MHz in, 28-30 MHz out
or 146-148 MHz with a second crystal
available for $5.95 each

A full description of this fantastic converter would fill this page, but you can take our word for it (or those of thousands of satisfied users) that it's the best. The reason is simple – we use three RCA dual gate MOSFETs, one bipolar, and 3 diodes in the best circuit ever. Still not convinced? Then send for our free catalog and get the full description, plus photos and even the *schematic*.

Can't wait? Then send us a postal money order for $49.95 and we'll rush the 407-A out to you. NOTE: The Model 407-A is also available in any frequency combination up to 450 MHz (some at higher prices) as listed in our catalog. New York City and State residents add local sales tax.

VANGUARD LABS

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250 MHz ELECTRONIC COUNTER

FREQ RANGE: 10 Hz to 250 MHz. **SENSITIVITY:** 100 mv from 2 MHz to 150 MHz, 200 mv from 150 MHz to 250 MHz. **RESOLUTION:** One-tenth Hz from 10 Hz to 10 MHz, 10 Hz from 10 MHz to 250 MHz. Has built-in prescaler for freqs above 10 MHz. **ACCURACY:** Two versions available, .0005% or .0002%, see below. **FEATURES:** Six digit incandescant readout with memory, two digit overrange for eight digit capability; six ranges; thru-line function measures transmitters up to 50 watts directly; self test; 1 MHz output jack; dual power supply, line or 12vdc for mobile use; polaroid filter; handle, bail, and cord holder. One year warranty. Prices are fob Reseda, Calif. Weight - 7 lbs.

Model 6-VHF-A with room temp xtal, .0005%, 15 to 30 degrees C $350.00

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AIL-124C Power Osc. 200-2500 MHz 365
Boonton 190A Q-MTR 20-260 MHz 375
HP100D-Freq. stand. w/scope-Acc. 1ppm 165
HP150A - 10 MHz Scope w/152B 185
HP160B(USM105)-14MHz dual trace scope 385
HP185B-DC-1gHz samp. scope w/186B 335
HP202B-Audio Gen.-5Hz-50kHz-.1-10v 75
HP212A-Pulse Gen. var. width and rate 65
HP330C-Dist. Anal. 20Hz-20kHz 225
HP522B-Freq. Counter-10Hz-120kHz-digital 45
HP540-B-Trans. Osc. for counter-to 12gHz 185
HP610B-Sig.Gen.-450-1200MHz, calib. attn. 365
HP686A-Sweep Gen.-8.2-12.4gHz 625
HP803A VHF Imp. bridge 50-500MHz 195
HP2590B-Precision Trans. osc-synchronizer 975
Kintel 301-DC standard-null voltmeter 220
NE 14-20C-Freq.-counter (sim. HP524C) 209
Nems Clark 1671-FM rcvr. 175-260MHz 210
Polarad TSA-spectrum analyzer 10MHz-44gHz (plug-ins available) 325
Polarad SA84W-spectrum analyzer, band switching 10MHz-41gHz 965
Rollin 30-Stand.sig.gen.40-400MHz-hi-pwr 585
Stoddart NM52A-RFI mtr, .375-1gHz, w/acc. 985
Tek RM15-DC-15MHz GP scope 295
ME26D/U-(HP410B) VTVM-to 700 MHz 85
SG24/TRM3 Sweep Gen. 15-400 MHz. CW, AM, FM Xtal markers, scope-Oev. to 20% 395
TS-403A-Sig. Gen. (HP616) 1.8-4gHz 385
URM-26 Stand. Sig. Gen. 3-400MHz 225
USM-16-Stand. Sig. Gen.10-440MHz AM-CW-FM-Pulse-Sweep, Phase-locked osc. 625

(Send SASE for complete list)

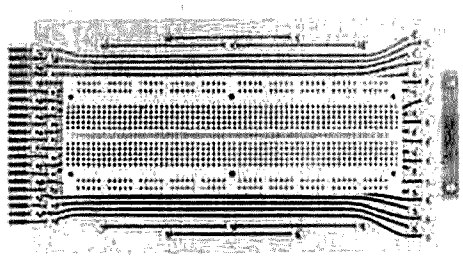
GRAY Electronics
P. O. Box 941, Monroe, MI 48161
Specializing in used test equipment

sub-audible tone encoder/decoder

Alpha Electronic Services has announced a new two-frequency subaudible tone encoder/decoder; the model DTSS-80 continuous-tone squelch system is exceptionally useful where a need exists to have a choice of tone frequencies for the purpose of controlling or selecting repeater stations, base stations or mobile radio units. By a simple two-position switch the dual tone makes possible the employment of multiple repeaters, increasing the range capabilities of a radio communications system. The selective calling of two base stations, two groups of mobiles or special control functions are other uses that this versatile device can accomplish. Each DTSS-80 tone unit is comprised of a plug-in encoder/decoder board and two plug-in TN-91 frequency determining boards. A special model is available that plugs directly into General Electric Master mobile radio units.

For information on the DTSS-80 write to Alpha Electronic Services, Inc., 8431 Monroe Avenue, Stanton, California 90680, or use *check-off* on page 94.

plug-in socket boards



EL Instruments announces the PL-1 and PL-2, Plug-in Socket Boards. The PL-1 and PL-2 feature one or two patented EL SK-10 component sockets on a 4½ x 9 inch glass epoxy G10 pc board.

The board includes a 22 pin card edge connector and card extractor handle. There are also 44 BP-23, breadboarding pins on the board for connecting the devices on the SK-10 sockets to the card

More Details? CHECK—OFF Page 94

edge connector. There is no soldering required; connections are made with standard 22 or 24 AWG hook-up wire.

The PL-1 and PL-2 are card rack compatible and have top edge test points for ease in circuit checking. The boards can be used to test prototype designs, add to existing card rack systems, update old pc card designs and many other bread-boarding functions.

The PL-1 (two SK-10 sockets), at $42.50 and the PL-2 (one SK-10 socket) at $29.95 are available from stock. Complete information and guantity prices are available from EL Instruments, Inc., 61 First Street, Derby, Connecticut 06418 or by using *check-off* on page 94.

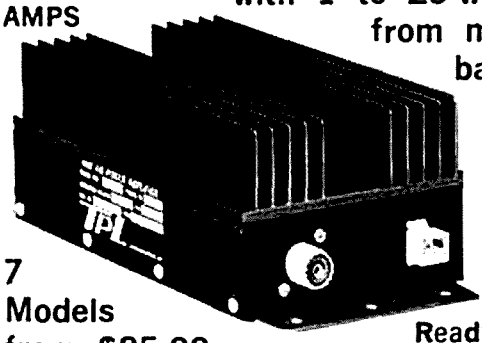
standard tone burst encoder

Standard Communications Corporation has introduced the new SC-ATBE-1 five-frequency tone burst encoder for activation of coded repeater systems. The unit offers exceptional stability of ±0.25% over a 180° temperature range through the use of a bridged-T oscillator and quality mica capacitors. The unit comes factory preset for the five most commonly used burst tone frequencies in the 1700 to 2500 Hz range – although the frequencies can be easily changed at any time.

The burst duration can be set from 0.1 to 1.0 seconds or keyed continuously by a front panel control for testing. Units come factory preset for 0.4 seconds. Output up to 15 mV rms, sine wave, is internally adjustable for different rigs with units coming preset for use with the SR-C826M. The unit styled to match the SR-C826M, mounts under the SR-C826M and plugs directly into the rear of any Standard Communications Corporation transceiver (except the hand-held units). The unit is powered by the standard 12 Vdc. It sells for $78.00.

More information is available from Standard Communications Corporation, 639 North Marine Avenue, Wilmington, California 90744 or by using *check-off* on page 94.

Get 'RIGHT ON' your REPEATER
with
CQC
2-Meter Crystals
TC-2M — .002% Tolerance
CRYSTEK crystals offer frequency stability over varying temperatures so you can go mobile in any environment.
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More Details? CHECK–OFF Page 94

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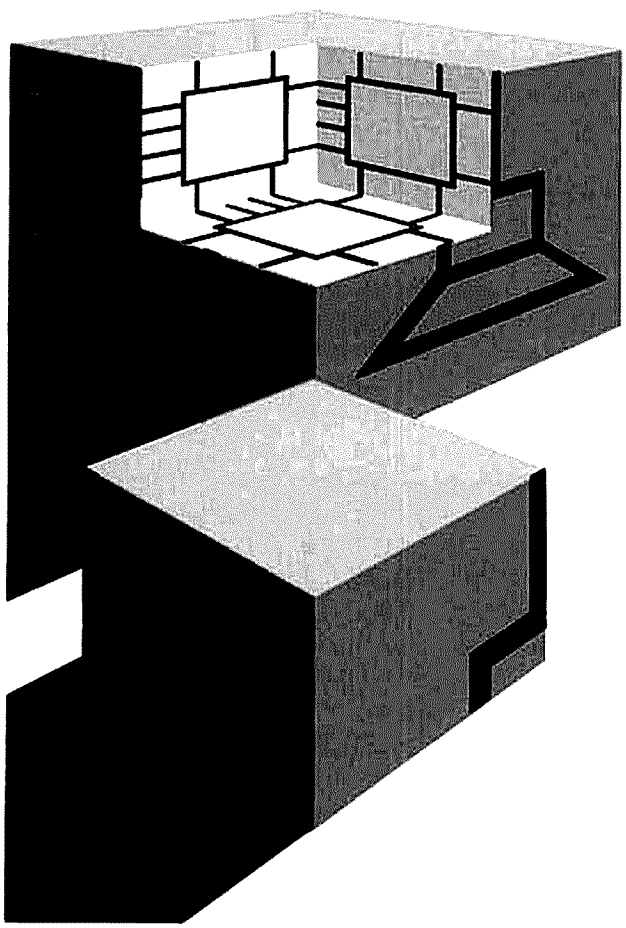
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ham radio magazine

MARCH 1973

solid-state 80-meter SSB TRANSCEIVER



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- AFSK generator 27
- electronic keyers 38
- mobile touch-tone 50

March, 1973
volume 6, number 3

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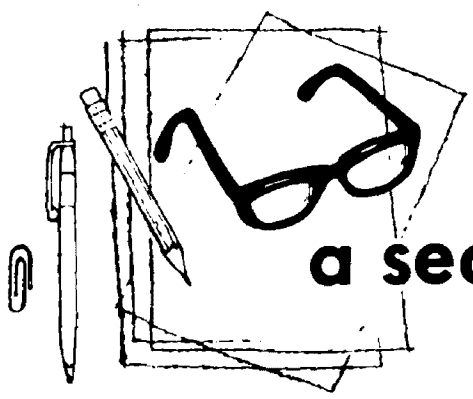
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a second look

by jim fisk

There are **so many nooks and crannies** to the hobby of amateur radio that it's difficult to say what facet is the most popular. Certainly, there is tremendous fm activity on two meters, and there's a host of operators up on the six-meter band, trying to add another state to their WAS list. Down on 75 meters there is a preponderence of rag chewers who gather on the same frequency, night after night. Then there is RTTY, slow-scan tv and traffic, brass pounders, county hunters and net nuts.

However, from listening on the air, I'd say that one amateur radio activity that is near the top of the list is certificate chasing and DXing. Actually, the two go together — what serious DXer do you know who doesn't have DXCC (as well as WAZ, WPX and at least a few others)?

With all the interest in certificates and awards, everybody and his brother is busy churning out another new one (with seals), for working continents, countries, counties, towns and club sites, all one mode and band seals, 25 cents extra. If you're looking for wallpaper, some of the certificates are worth applying for, but all too many times they are poorly printed on a lousy grade of paper and don't even warrant space in your round file.

It's been my experience, as a one-time certificate chaser, that operating awards offered by national amateur radio societies (ARRL, RSGB and NZART, for example) are well done; awards offered by national magazines are usually worthwhile as are the beautiful awards sponsored by the YL International SSBers. But, for every nice certificate available, there are a dozen others that would make passable toiletries.

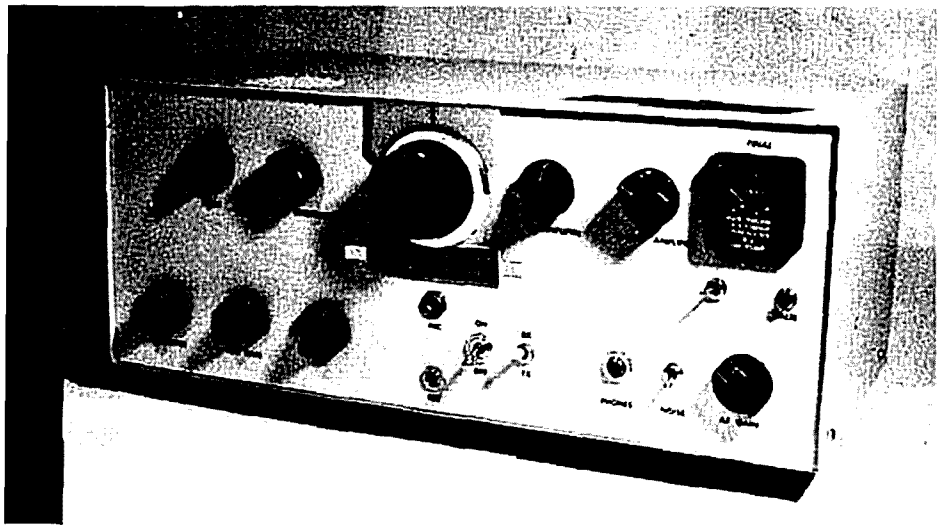
You can usually predict the type of certificate you're going to get by return mail by considering the sponsor, the difficulty of the award *and* the cost. If the award is for working three members of the Podunk Amateur Radio Club while on safari to Omallabug county, and they want 50 cents to cover postage and handling, don't expect too much! On the other hand, when the Organization of American States Association offers an award for working all member nations (WAAN) at no cost, you can look forward to a handsome certificate.

It is unfortunate that the biggest bulk of junk certificates seems to originate in the United States. The certificates from overseas are almost always very tastefully done and are a welcome addition to the hamshack wall. I think it's high time we brought some of our homegrown awards up to snuff.

If your club offers a certificate of any kind, get it out and take a good, close, unbiased look at it. First of all, is it printed on a good grade of paper? (It doesn't have to be on parchment, but the paper shouldn't look like it escaped from a newsprint factory, either.) How about the printing? Are the letters clear and sharp? Are there ink smudges and dirty fingerprints, deposited by a careless printer? Finally, was your name and callsign scribbled on the certificate by some refugee from the third grade, or is it carefully lettered or typewritten?

If your club award passes these three simple tests, congratulations! Put it back in its frame and hang it on the wall. If it doesn't pass, resolve to take it up with your fellow club members at the next meeting. Let's relegate all those junk certificates to the trash can.

Jim Fisk, W1DTY
editor



solid-state 80-meter ssb transceiver

Complete circuit details for an all solid-state 10-watt ssb transceiver — building-block construction simplifies future circuit revisions

Dave Hembling, VE6ABX, Box 728, Pincher Creek, Alberta, Canada

This single-band 80-meter ssb transceiver was built using a combination of solid-state building-blocks in crucial circuitry, as well as descrete components in subordinate circuits. No transmit-receive switch, as such, was used; instead, seven tiny surplus hermetically-sealed relays were used and located at the site of each associated building-block. Also, the number of tuned circuits was kept to an absolute minimum. For example, the only tuned circuits employed in the receive mode are in the front-end, the vfo and optionally in the i-f amplifier.

Plans for revision will eventually eliminate the tuned circuit in the i-f amplifier stage when the present MPF102 fet and single tuned circuit are replaced with a single broadband IC amplifier (for example, the MC1350P gives 60 dB gain at 60 MHz and is available for under two dollars).

In the transmit mode, the only tuned circuits are in the vfo and the buffer, and the driver/final. The only tuning needed on transmit is the vfo and the collector of the final. The three tuned circuits ahead of the final are staggered-tuned to give reasonably broadband performance across the 200 kHz I use most. There is plenty of drive, as the mike gain control is about

half open for the beginning of clipping.

One of the advantages of the building-block method of construction is that it allows for constant and unpredictable later modifications and changes in the circuit as various other ICs and discrete components become available. For my part, about 90% of the enjoyment of amateur radio is found at the workbench, and this little rig is built to provide for an almost never-ending series of re-engineering.

The heart of the rig is a Snelgrove* F9000-1 crystal lattice filter with a bandpass of 400 Hz to 2500 Hz at 6 dB down. Filter insertion loss is only 1.5 dB and 50 dB suppression of the unwanted sideband is provided.

The other most significant block used in the design is the Motorola MC1496 integrated-circuit dual-differential amplifier which offers exceptionally attractive characteristics as a balanced modulator, double-balanced mixer and product detector. As a balanced modulator, when

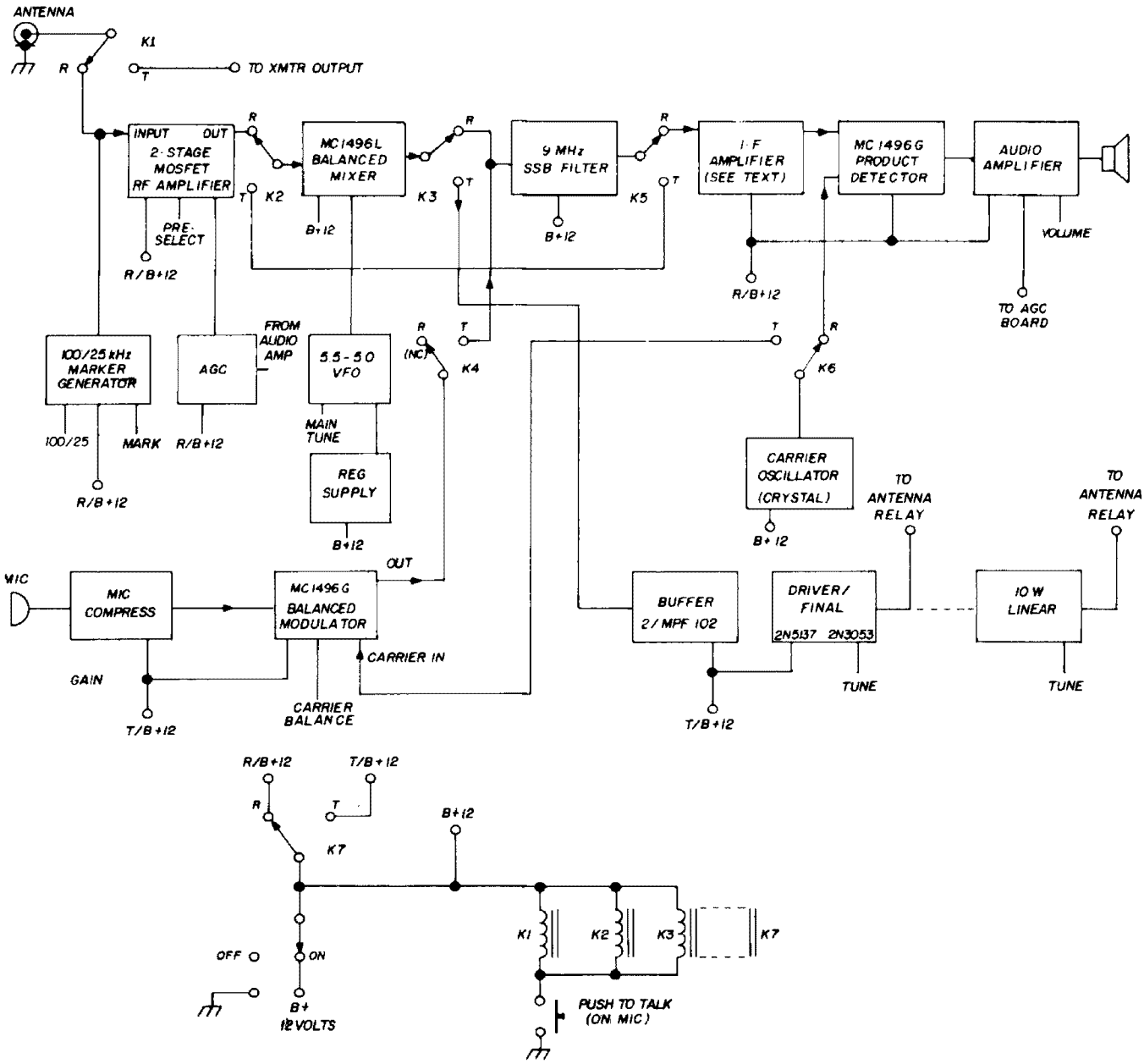


fig. 1. Block diagram of the 80-meter ssb transceiver. Relays are used for the bulk of the transmit-receive switching.

*C.R. Snelgrove Company, Ltd., 141 Bond Avenue, Don Mills 404, Ontario, Canada.

supplied with the appropriate signal levels, the MC1496 easily provides 35 dB carrier suppression, 50 dB spurious sideband suppression and 20 dB suppression of the second carrier harmonic. These figures, along with the 50 dB suppression of the crystal filter, add up to a very respectable single-sideband signal at the antenna. Also, the MC1496 provides for carrier balance adjustment through a dc potentiometer, and simplifies things considerably in this respect.

It should be noted that the MC1596, which was written up by K7QWR,[1] is almost the same as the MC1496, and this tremendous little IC is used to good advantage in this rig in three key circuits. The most significant difference between the MC1596 and MC1496 is in their respective operating temperature ranges. The 15-series meets exacting military requirements (-55 to +125 degrees C), while the 14-series is designed for operation in the 0 to +70 degree C range.

The MC1496 offers only slightly less carrier suppression than the MC1596 (see the specification sheet for the MC1596/1496 as well as Motorola's very helpful Application Note AN-531). The MC1496, therefore, is entirely adequate for most amateur applications. Also, the 14-series is approximately half the cost of the 15-series device, another worthwhile consideration.

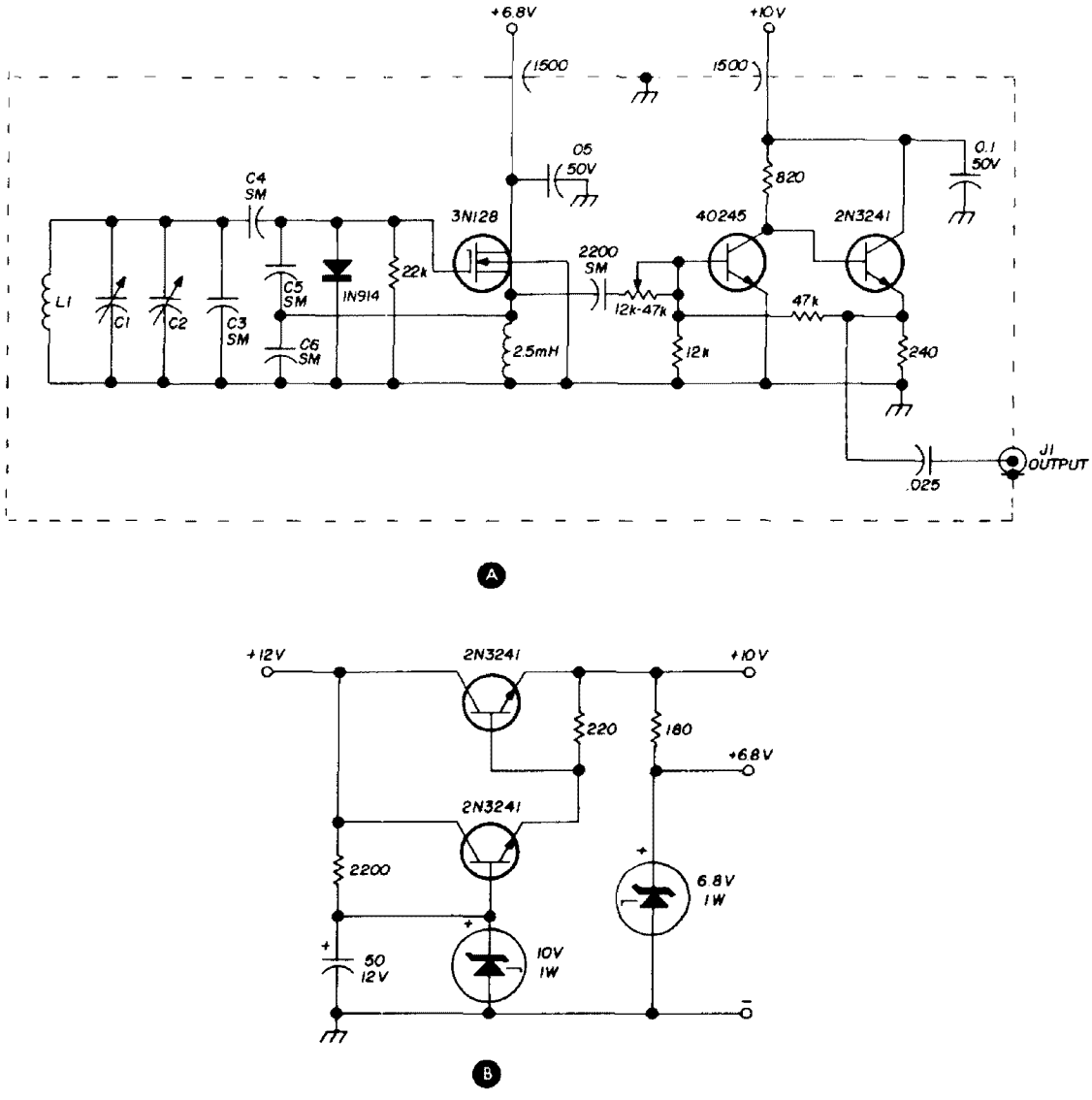


fig. 2. High-stability mosfet vfo tunes from 5.0 to 5.5 MHz. Regulated power supply is shown in (B).

basic interconnections

The two blocks that make this rig such a good performer are the high quality filter, with its nice steep skirts, and the

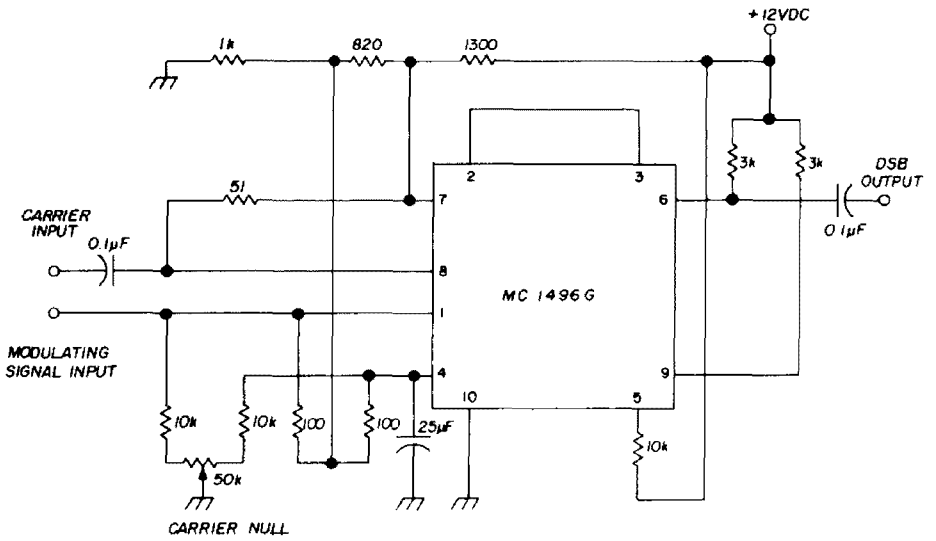


fig. 3. High-performance balanced modulator circuit uses Motorola MC1596 or MC1496 IC.

MC1496 ICs. In order to achieve maximum performance from the filter, it must be impedance matched at both input and output. The circuit shown in fig. 7 meets the manufacturer's requirements, although the F9000-1 specifications were originally designed for use around vacuum-tube circuitry. The circuit in fig. 7 uses two MPF102 field-effect transistors and provides a perfect match for the filter.

For maximum performance from the MC1496, however, the crucial variables rest in the voltage levels supplied to the differential inputs. The MC1496 has an excellent dynamic range, as demonstrated by the 90 dB figure which it provides when used as a product detector. However, to achieve maximum carrier suppression as a balanced modulator, the levels of carrier and audio supplied at the inputs must be carefully set to within prescribed limits. Recommended signal levels for the MC1496, under the three utilized functions, are given in table 1. For obvious reasons no figure is given for the ssb signal levels. Once the given voltages are obtained, however, it is a relatively straightforward matter to experimentally adjust coupling parameters to achieve the desired operating characteristics.

For example, I found that a small amount of i-f gain was needed to achieve good product detector operation. The original plan was to operate with an i-f gain of unity, but this resulted in inadequate ssb signal levels for good mixing in the MC1496. Very little additional signal was required to obtain the desired results from the product detector.

table 1. Recommended signal levels for the MC1496.

function	input at pin 8	input at pin 1
balanced modulator	carrier oscillator, 60 mV rms	audio, 3 - 300 mV rms
double balanced mixer	vfo 100 mV rms	ssb signal
product detector	carrier oscillator, 300 mV rms	ssb signal

the vfo

The only change from W2YM's original plans, apart from greatly reduced enclosure size, was the coil/tuning capacitor combination. I used a ceramic, slug-tuned surplus coil form and have found it to offer very slight upward frequency drift

without temperature compensation. However, a few picofarads in parallel with a +650 temperature coefficient served to stabilize the drift. This determination is best done with a digital counter, over a

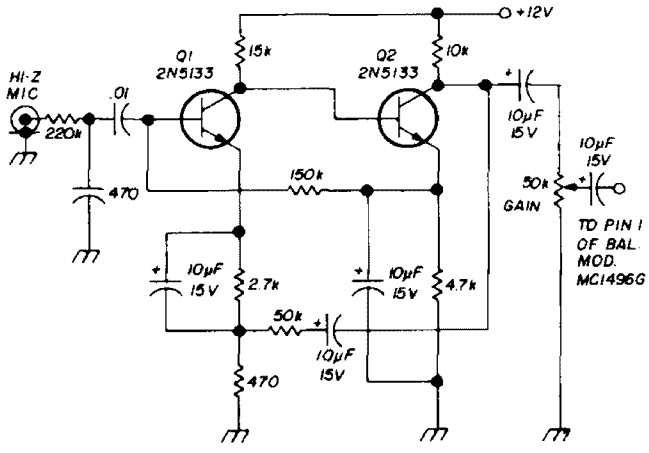


fig. 4. Simple speech compressor circuit designed by VE6BB.

period of several days. The value needed depends upon the physical characteristics of the inductor you use, but this is open to considerable latitude of design. My vfo was designed to give a 500-kHz tuning range, from 5.5 to 5.0 MHz, but the vfo could just as easily be designed to cover a smaller range if access to the whole band is not desired.

One of the later modifications I'm keeping in mind is the possibility of putting the rig on 20 meters, which would require the full 500-kHz vfo spread. (i.e., 9.0 MHz minus 5.0 MHz = 4.0 MHz; 9.0 MHz minus 5.5 = 3.5 MHz, and 9.0 MHz plus 5.0 MHz = 14.0 MHz; 9.0 MHz plus 5.5 MHz = 14.5 MHz.) To put the rig on 40 meters would require an extra converter stage.

balanced modulator

The Motorola MC1596G IC was used as the balanced modulator because the MC1496 was not commercially available at the time I built the circuit shown in fig. 3. The speech compressor and balanced modulator circuits were etched on a small printed-circuit board 2½-inches square. Care was taken to provide shielding between this module and all others, particularly the carrier oscillator.

Both the balanced modulator and carrier oscillator were shielded on all six sides. Rf chokes and feedthrough capacitors were placed in all B+ and relay leads in and out of the shielded compartment. Carrier balance and microphone gain controls were brought out to the front panel with leads no more than 0.75-inch long.

The speech compressor circuit, fig. 4, is built on the same PC board as the balanced modulator. This ingenious little circuit was designed by Basil Barnes, VE6BB, and works very well. The B+ supply to the speech compressor is isolated from the balanced modulator with a series-connected 280-ohm resistor and 0.1-µF bypass capacitor.

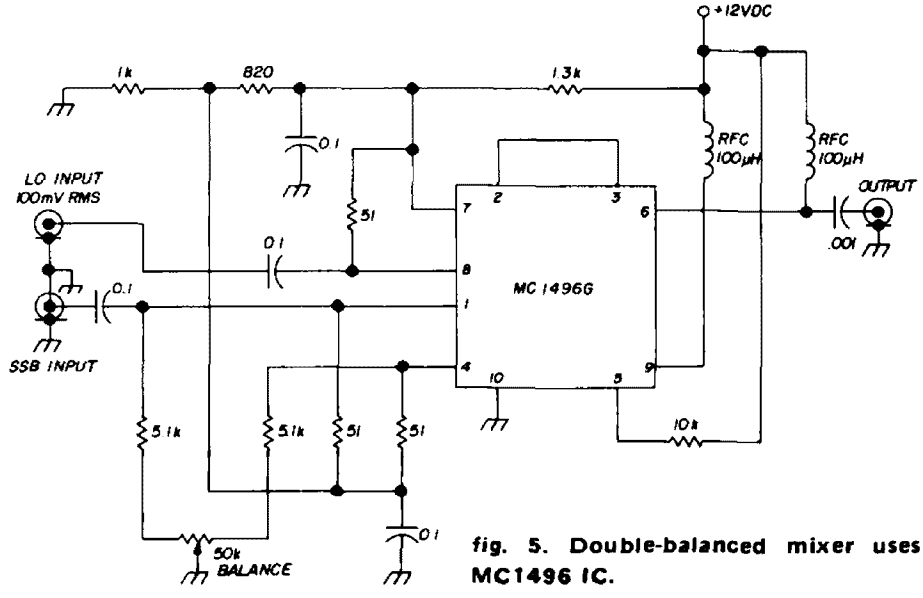


fig. 5. Double-balanced mixer uses MC1496 IC.

All parts for the double-balanced mixer shown in **fig. 5**, including the MC1496G IC were mounted on a small PC board. The board was mounted on the chassis so the 50k balance adjustment

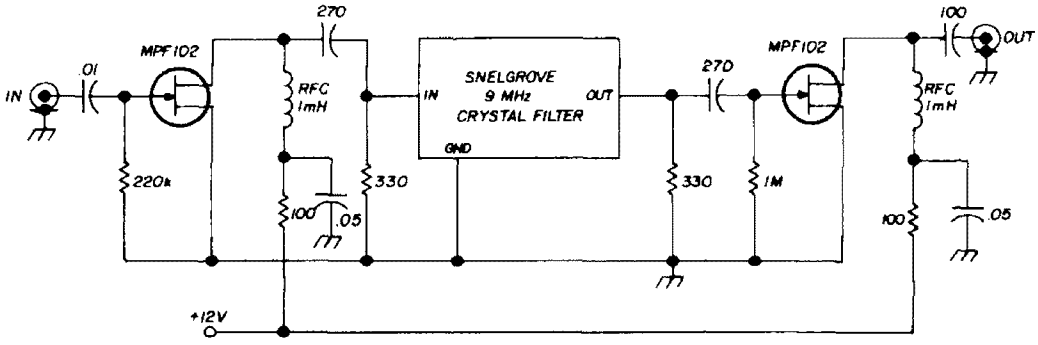


fig. 7. Field-effect transistors are used to match the input and output impedances of the Snelgrove 9-MHz crystal filter. The 1-mH rf chokes are Hammond miniature powdered iron core types; capacitors are 50-volt ceramic discs.

potentiometer could be reached through a ¼-inch access hole in the chassis.

The product detector (see **fig. 6**) needs no adjustment and was built on a small PC board which was mounted in a Minibox installed on top of the main chassis. Carrier injection to the product detector is through a 52-pF capacitor. Signal output to the audio stage is through a 0.47-μF disc capacitor. (As with the other building blocks in the rig, the value of the coupling capacitor must be determined experimentally for best performance.)

crystal filter

Two MPF102 fets were used to make up the 1.5 dB insertion loss to the filter and the overall circuit has approximately unity gain (see **fig. 7**). The only caution is that it is necessary for good suppression that the input electronics not be able to "see" the output electronics, except through the window of the filter. Any stray coupling between the input and output circuitry will undermine the filter's suppression capability. The filter itself should be mounted so that its metal can presents an rf barrier to the two MPF102s and their associated circuitry. The PC board was made extra large to assure good isolation between the input and the output of the 9-MHz filter.

The carrier oscillator circuit (**fig. 8**) uses the Snelgrove crystal, and the circuit should be provided to the Snelgrove company when the filter and crystals are ordered. Some carrier level adjustment is

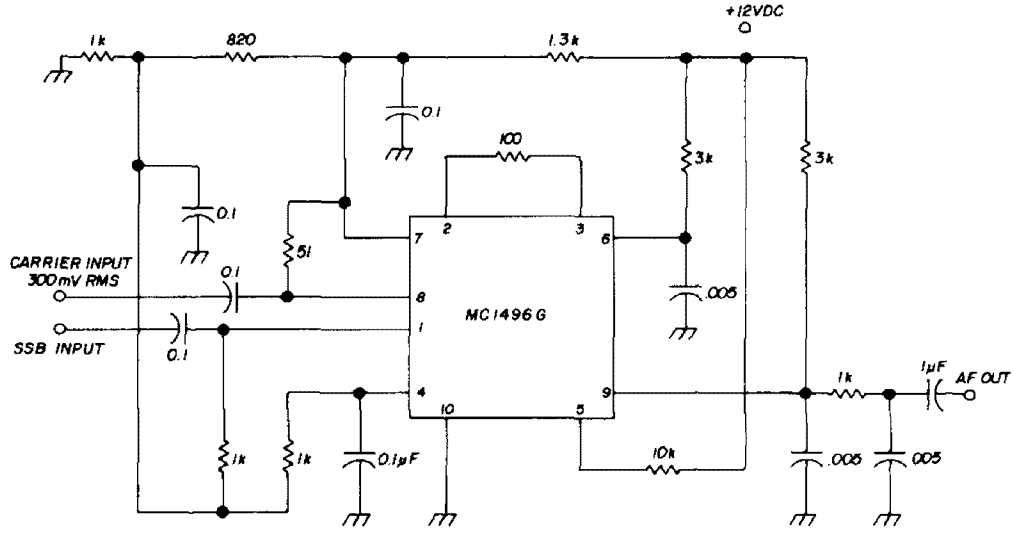


fig. 6. Product detector circuit.

available by adjusting the iron-core, slug-tuned coil, but better adjustment is facilitated by the 30k trim-pot. The trim-pot should be accessible through a hole in the chassis when the shielded

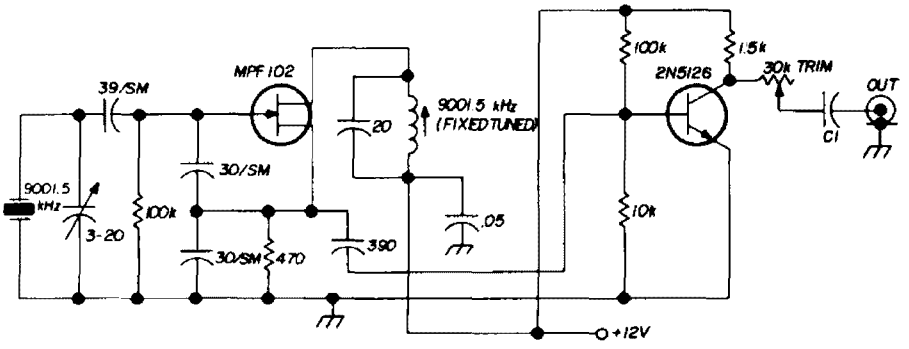


fig. 8. Circuit for the crystal-controlled carrier oscillator. The value of C1, 50 to 330 pF, is chosen for desired output range.

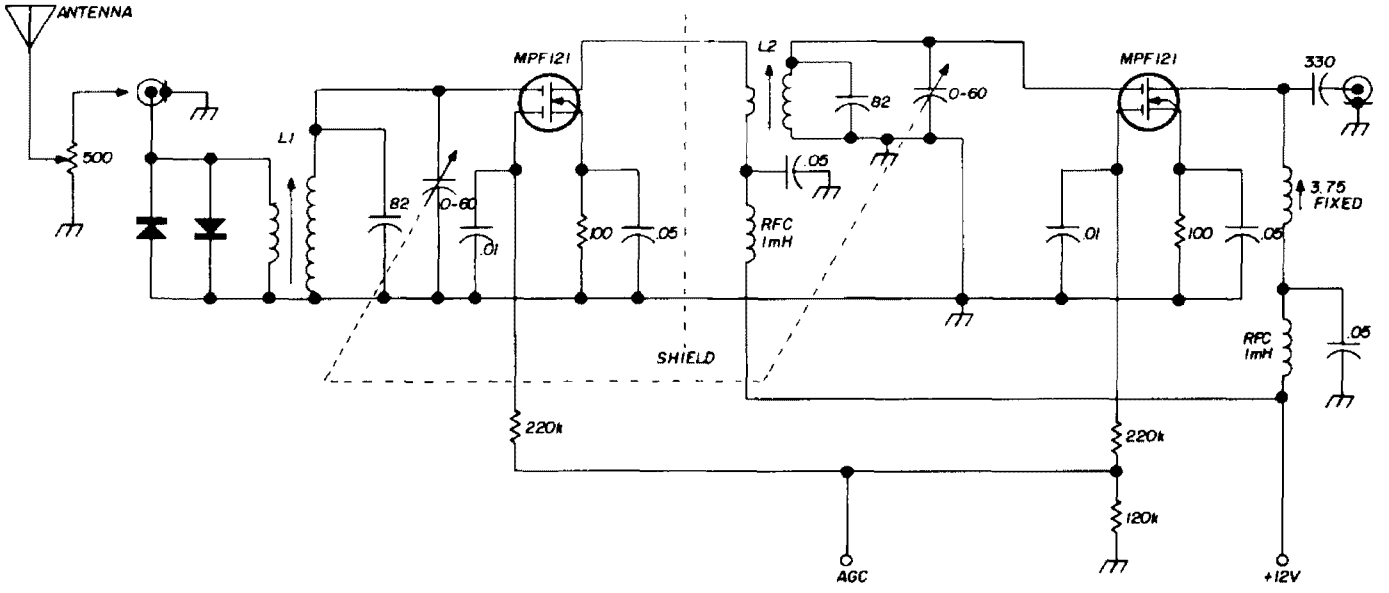
compartment is closed up. Rigidity of construction is just as important for this circuit as for the vfo if stable operation is to be obtained.

receiver

A two-stage mosfet rf amplifier was used in the receiver (see fig. 9). Receive sensitivity was determined only by comparing overall performance to the communications receiver I normally use, a Drake R4B. The noise figure of the homebrew receiver was not quite as good as the R4B, but sensitivity seems to be nearly similar. The agc operation of my receiver doesn't compare too well with the R4B, but this is probably the weakest point in the receiver's operating characteristics.

The mosfet front end shown in fig. 9 performs about twenty times better than the MPF102 fet I tried earlier. Also, better agc performance is available with the MPF121 mosfets.

The front end is tuned with a dual



L1 40 turns no. 30 on ¼" slug-tuned coil form. Antenna winding is 10 turns no. 25 on ground end of same form. Windings are isolated by 3/8" length of plastic sleeving

L2 40 turns no. 30 on ¼" slug-tuned coil form. Input link is 15 turns no. 30 on ground end of gate coil, two windings isolated as with L1

fig. 9. Receiver rf amplifier uses two MPF121 mosfets. The two back-to-back diodes across the antenna terminals are small-signal silicon switching types. Rf gain control is simple 500-ohm potentiometer voltage divider in the antenna lead.

365-pF broadcast variable (remove all but two plates per section). The preselector adjustment offers good selectivity, and the front end is very responsive to a resonant antenna. The simple rf gain control, a 500-ohm potentiometer voltage divider between the antenna and the input to the first stage, works well.

At the present time I am using the simple hang agc circuit shown in **fig. 10.** The audio input is picked up from the audio stage prior to the volume control. A 2-megohm trim-pot provides agc voltage control.

A superior agc system is illustrated in **fig. 11.** This agc system, originally described by DL6WD[3] uses a single RCA CA3035 IC. With this agc system the input signal is taken from the unused differential output of the MC1496 product detector at pin 6 through an appropriate coupling capacitor. With this system, the discrete audio output stage can be replaced by an IC such as the Motorola MFC9020 2-watt audio IC for superior audio performance.

either 25- or 100 kHz markers. A momentary-contact pushbutton, **S1**, activates the circuit by completing the source circuit of the fet. Two inexpensive μL923 J-K flip-flops were used to divide the 100-kHz

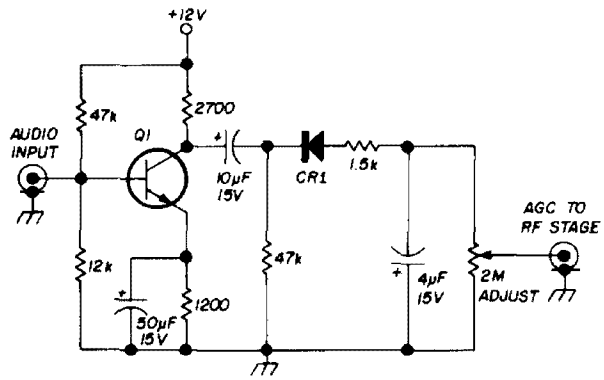


fig. 10. Simple hang agc circuit. Audio input must be picked up before the volume control (see fig. 15). Transistor Q1 is 2N5137, 2N2222, etc. CR1 is any small-signal silicon switching diode.

crystal frequency. The 20-pF trimmer is adjusted to zero beat the 100-kHz crystal signal against a receiver tuned to WWV.

i-f amplifier

As I mentioned earlier, the original

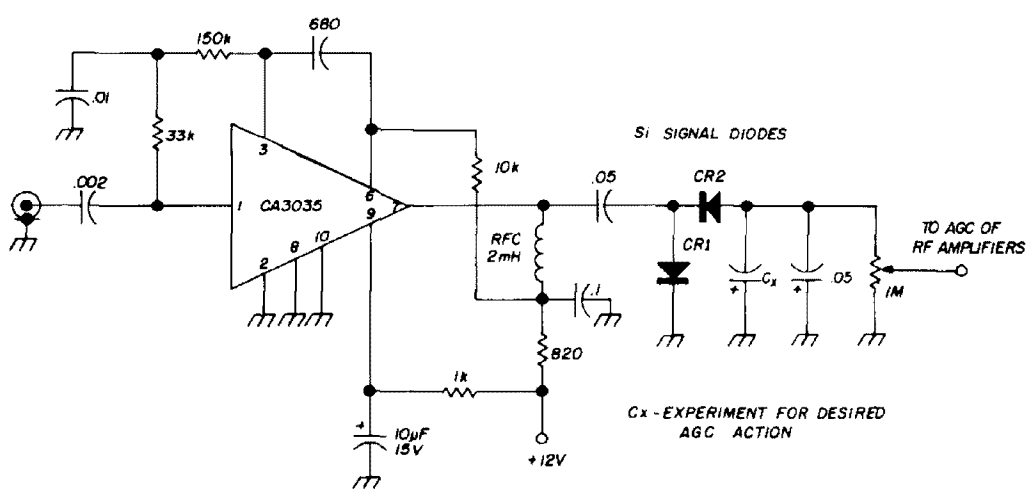


fig. 11. Integrated-circuit agc system. CR1 and CR2 are silicon signal diodes. The value of Cx is chosen for desired agc action.

marker generator

The circuit for a very useful crystal-controlled frequency marker is shown in fig. **12.** This circuit, which is similar to the circuit used in the Drake R4B receiver, provides front-panel control of

plan was to use an effective i-f gain of unity. However, I found that product detector performance was enhanced tremendously by adding a small amount of i-f gain. A two-stage fet i-f stage with one 9-MHz tuned circuit in the input gate and 1-mH rf chokes in both drains gave better

performance than the circuit shown in **fig. 13**, which I used, but due to space considerations this postage-stamp sized circuit was installed until such time as an IC can be put in its place.

16. Depending on the frequency range which is used, these two stages, as well as the pre-driver buffer in the next circuit block, can be stagger tuned, leaving only the collector of the final power amplifier

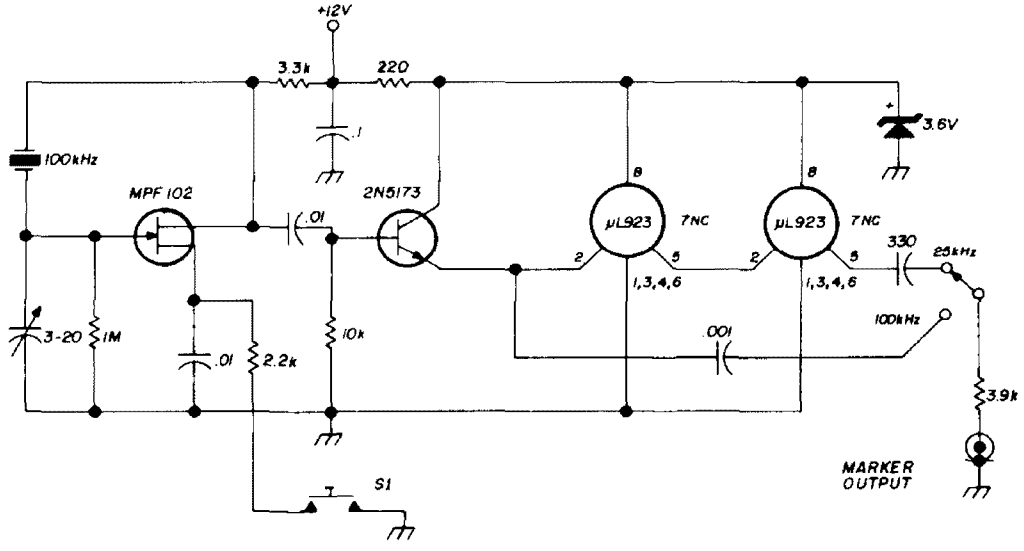


fig. 12. Crystal marker generator provides 25- and 100-kHz markers.

A good candidate for this job is the Motorola MC1350P, which, without any tuned circuits, can provide 35 dB gain. The MC1350P i-f circuit shown in fig. **14** was found to offer excellent performance with the presettable gain control connected to the agc input of the IC. It would be difficult to incorporate the device's agc input into the receiver agc system due to the low impedance of the IC. However, i-f stage agc would not offer any special merit anyway, because the MC1350P does not add seriously to receiver noise at the moderate gain level at which it is used.

It is hardly necessary to include data on the audio output amplifier (fig. **15**) except to mention that it is handy to have one of the speaker leads at ground potential, not always a feature of IC amplifiers. With a grounded speaker system, a two-circuit headphone jack can be used to switch the output.

transmitter

A simple two-stage fixed tuned amplifier with two MPF102 fets is used for the transmitter buffer circuit shown in fig. which must be tuned from the front panel.

Alternately, if access to the entire 80-meter band is desired, the tuned circuit may be tuned with a ganged variable capacitor which is brought out through the front panel. Since the transmitter buffer has relatively high gain, care must be taken to isolate the input from the output.

There have been a number of solid-state transmitting circuits published in the past, but most have been designed for low-power CW, so they are not suitable for linear ssb operation. When working with the circuit shown in fig. **17** I gained

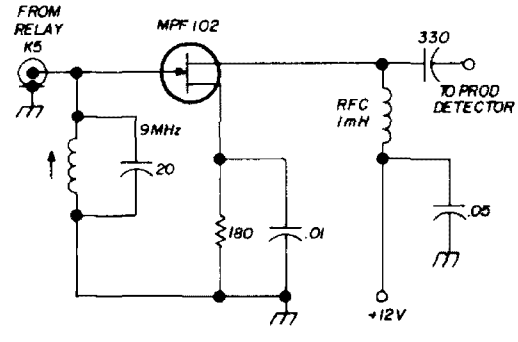


fig. 13. Simple one-stage, unity-gain 9-MHz i-f amplifier uses one fixed-tuned circuit.

considerable insight from W3TLN's experience with biasing QRP transistor finals to obtain linear operation.[4] Other articles contributed to the circuit I eventually used, but since base bias current is

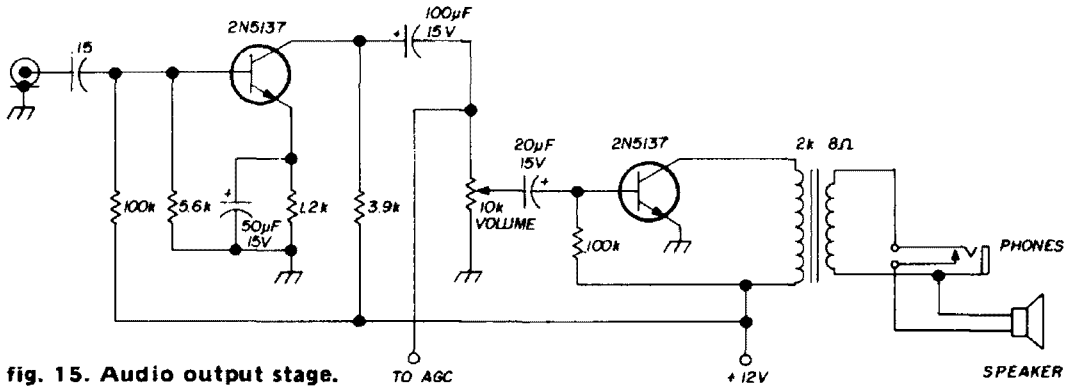


fig. 15. Audio output stage.

the crucial factor when using transistors in a linear amplifier, my own improvement was to add zener regulation to the bias voltage supply.

The 36-volt zener in the collector circuit clips any peaks beyond that voltage. The value of the capacitor, C1, is adjusted experimentally to resonate L1 at the desired center-band frequency, and falls in the range from 100 to 330 pF. The 20k trim-pot is adjusted for 5 to 8 mA of idling collector current.

Later, I added a 10-watt linear to the rig which is easily driven by the 1.5-watt stage. The circuit is shown in fig. 18. This

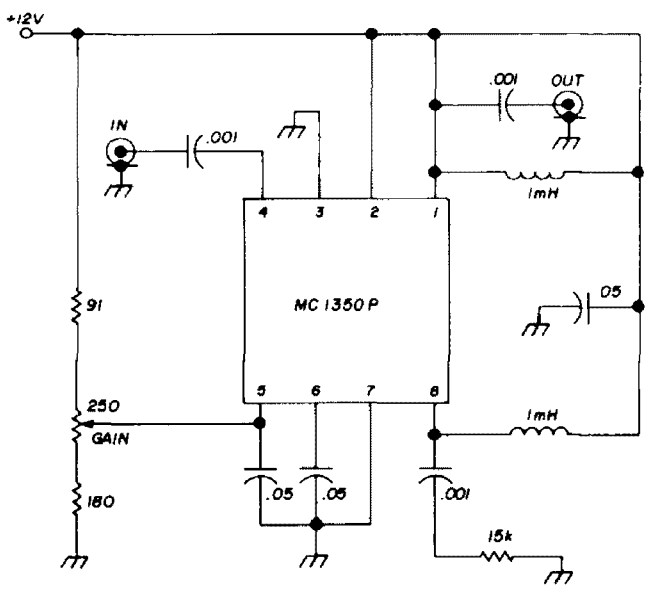


fig. 14. IC i-f amplifier which is suggested by the author for somewhat better performance.

circuit uses a Fairchild SE9081 which has a power limitation of 42 watts and a frequency cutoff of 70 MHz. Cost of the SE9081 is less than $2.00.

The 2-inch heatsink of the SE9081 is mounted on 1-inch ceramic standoffs with a small PC board mounted on the other end of the standoffs. The SE9081 could easily be driven to 2 to 2.5 amps of collector current (24 to 30 watts input), but running the device at a cool 10 watts input provides a good safety factor so the rig can be operated safely for a moment or two with a mismatched antenna load. The 20k trim-pot is adjusted for 8 mA of idling collector current.

power supply

For portable operation, I use a 7 amp/hour rechargeable lead-acid motorcycle battery. This provides a very stable supply voltage over extended operating conditions. For mobile operating I plug a cord into the car's cigarette lighter. The rig requires about 150 mA on receive, and 1.5 amps on transmit. The high transmit current drain is due, in part, to the number of 12-volt relays which are used for the transmit-receive switching.

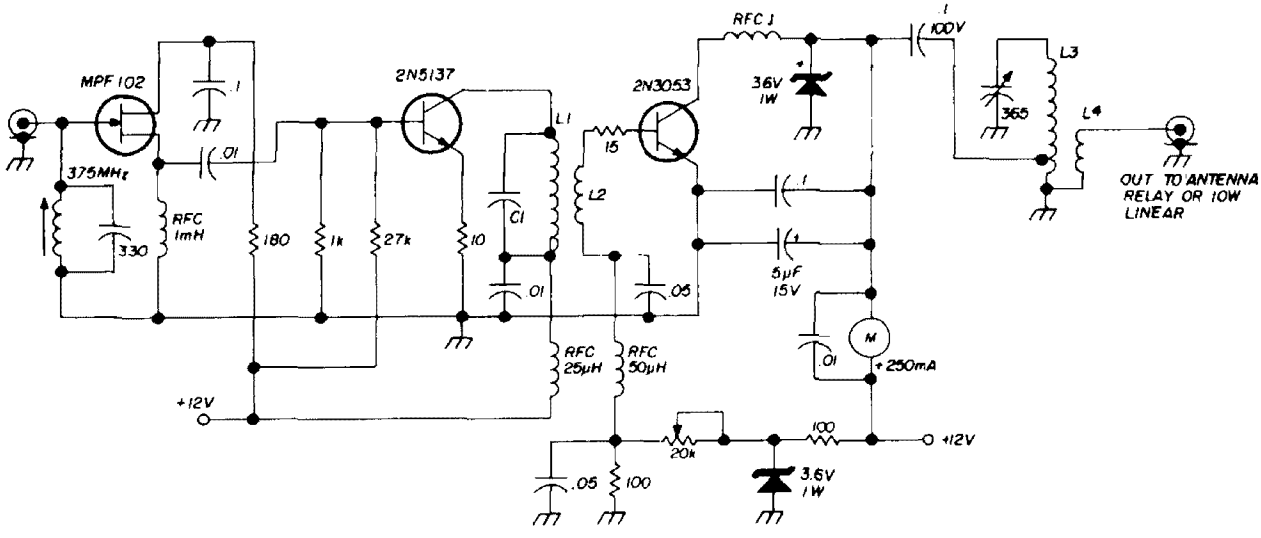
construction techniques

From my point of view, the average experienced amateur homebrewer is not looking for Heathkit-style plans before undertaking a new project. Physical dimensions and chassis layouts are, in my view, quite unnecessary for the average home-brewer. The only crucial data are the details on the electronics, and those

physical matters which affect electronic functioning, such as shielding.

The circuits used for this rig were adapted and borrowed from a variety of sources. Usually, however, only the cable and the blocks adjacent electronically may be located at opposite ends of the enclosure physically, provided adequate shielding is used.

Stability, of course, requires that the



L1	25 turns no. 24 enamelled on Amidon T-68-2 toroid core. (Use entire core surface)
L2	7 turns no. 24 enamelled, spaced evenly over L1
L3	7 mH. 34 turns no. 24 wound on Amidon T-68-2 toroid core, tapped 4 turns from ground end
L4	6 turns no. 22 insulated hookup wire wound over tapped portion of L3
RFC1	3 Amidon ferrite beads on 3/8" length of no. 22 wire mounted as close as possible to the transistor
RFC2	25 mH rf choke (James Millen J-300-25)

fig. 17. Circuit for the driver and 1.5 watt final (linear) amplifier. The 2N3053 power transistor requires a small heat sink for proper cooling. The 20k trim-pot sets collector idle current so the transistor operates as a linear amplifier.

schematic was utilized, and the physical form of construction was determined solely by electronic requirements and available materials. For example, the vfo, which is electronically almost exactly the one designed by W2YM,[2] was built in a small metal box and works perfectly. The original plan calls for a much larger enclosure with front panel measuring 7 by 10 inches.

The rig was built around a Hammond aluminum chassis (3 x 8 x 16 inches). The front panel is 16-inches wide by 6¼-inches high. Actually, the whole rig could be built in about 2/3 this space, and for the sub-miniature minded, this should be kept in mind. But then, layout is non-critical when using the modular building-blocks technique, as each block is interconnected by RG-174/U coaxial

vfo and carrier oscillator be constructed as rigidly as possible so that physical stress on the cabinet itself produces the minimum corresponding frequency change.

All circuits were built on single-sided

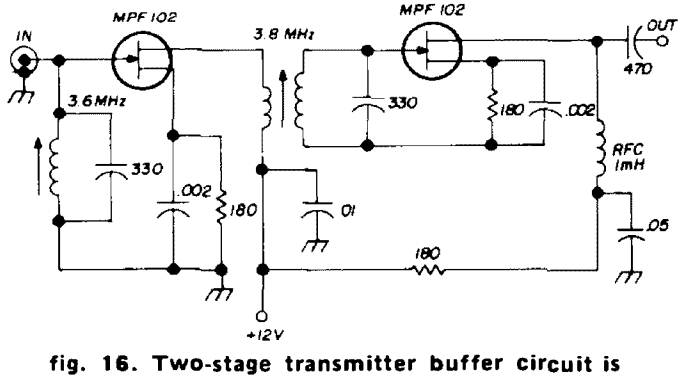


fig. 16. Two-stage transmitter buffer circuit is stagger tuned to cover the Canadian phone band (first stage tuned to 3.6 MHz, second stage to 3.8 MHz).

epoxy printed-circuit boards, and these are mounted by one or two threaded metal standoffs. It would be possible to build the entire rig on only several larger boards. However, the point of the building-block approach is to permit maximum flexibility for later changes and PC boards do not lend themselves to later modification, except in the sense of replacement of the board itself.

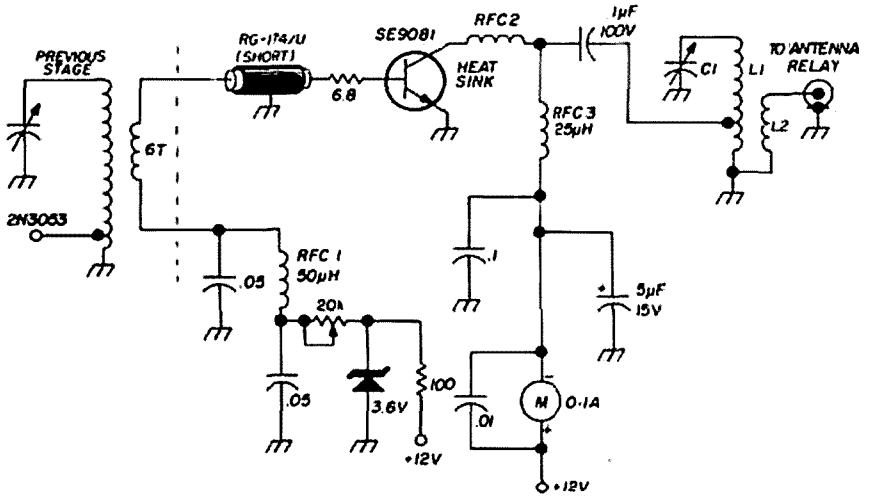


fig. 18. Ten-watt linear amplifier. Idle current is set by 20k trim-pot. RFC3 is home-made, to carry 1 amp; see text.

C1 165-pF variable in parallel with 100- to 330-pF fixed ceramic

L1 34 turns no. 19 on Amidon T-80-2 toroid core, tapped 4 turns from ground. Antenna link is 6 turns no. 22 insulated hookup wire, wound over 4-turn tapped section

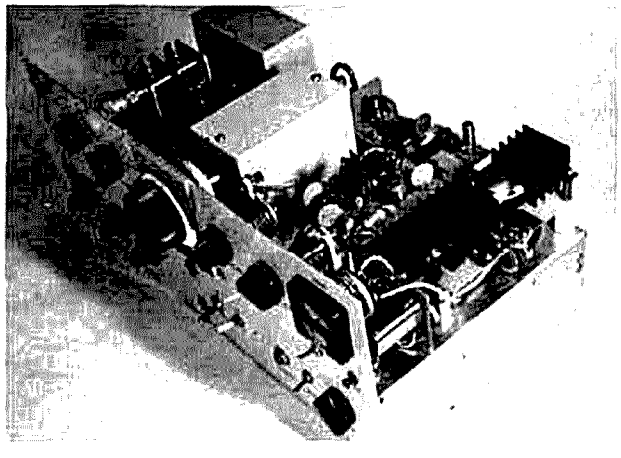
RFC1 94 turns no. 31 on Amidon T-50-2 toroid core

RFC2 2 ferrite beads on no. 19 wire, close to SE9081 socket

RFC3 67 turns no. 22 enamelled on Amidon T-50-2 toroid core

test equipment

A vtvm with rf probe is required to determine appropriate signal levels between the various blocks of the rig. The other necessary tool is a good quality general-coverage receiver (ssb) capable of tuning the 5.5 to 5.0 MHz and 8998.5 to 9001.5 kHz range. The more accurate the receiver, the more precisely carrier suppression can be set, and desired filter action obtained. Of course, the use of a digital frequency counter makes things a lot easier, especially when constructing the vfo. The vfo could be set up by the use of a general-coverage receiver, but there would be two drawbacks to this: first, the resultant vfo calibrations would be limited to the accuracy of the receiver, and secondly, it would be impossible to determine and correct drift problems.

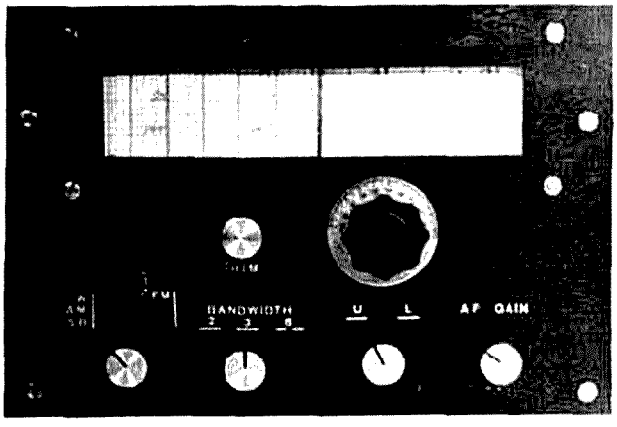


Layout of the 80-meter ssb transceiver.

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ham radio



all-mode companion receiver

The reciprocating detector makes its debut as an fm discriminator in this receiver design

Stirling M. Olberg, W1SNN, Waltham, Massachusetts 02154

When I encouraged my wife and daughter to become hams, they took over my receiver and transmitter, which were equipped with outboard converters for six and two meters. This situation left me without means for chasing DX on 20 meters, so some changes were in order. The all-mode companion receiver described here was designed and built for use with vhf converters so I could retrieve my receiver, an R4A, for DX work.

features

The all-mode companion receiver uses solid-state devices available on the surplus market. Most are available from advertisers in the amateur magazines. Construction is not difficult for amateurs who like to build their own equipment. Substitution of ICs and diodes can be made easily. All transistors should be npn silicon devices that work up to 50 MHz. The fets, however, should be those shown, which are also available from surplus sources.

The receiver uses a reciprocating detector.[1] This circuit works extremely well as an fm discriminator and as a synchro-

nous detector for a-m and CW. In a nonsynchronous mode, it's an excellent detector for ssb. A narrow filter in the circuit helps provide impulse noise suppression.

design development

The two converters in use at my station require an input frequency of 14-18 MHz for their i-f strips, so the first mixer operates within this range (fig. 1). The second converter input is 1.5 MHz and output is 500 kHz. Why 500 kHz? That's easy. I had three mechanical filters designed for a 51J4 — 0.2 Hz, 2.8 kHz and 6 kHz were the bandwidths — nice for CW/ssb and maybe a little sharp for fm, but okay for a-m, so these goodies were included in the design.

The question of signal leakage from coastal and maritime stations, which use 500 kHz as a calling frequency, was resolved by comparing the completed all-mode companion receiver with three other very fine receivers that use 500-kHz i-fs: a Collins 51J4, a 51S1, and a military version of the 51J4 known as an R-388/URR. The three Collins receivers were connected to a common antenna along with the all-mode job and tuned to 14 MHz. A BC453 receiver was tuned up on 500 kHz and WCC, WSL, and many other coastal stations were heard very strongly. Not a trace of these signals was detected on any of the four receivers, which were operating simultaneously.

A comparison of the internal shielding of the three Collins receivers indicated almost identical construction. Lead dress and bottom plates were arranged to inhibit coupling of external signals. The all-mode set uses quite a bit of decoupling and extremely tight shielding, which accounts for its very good rejection of signals on 500 kHz. The first conversion i-f at 1.5 MHz performs just as well for the same reasons. If the coax cables described are used to couple one unit to the other, and good shielding is used, no problem should be encountered with feed-through interference.

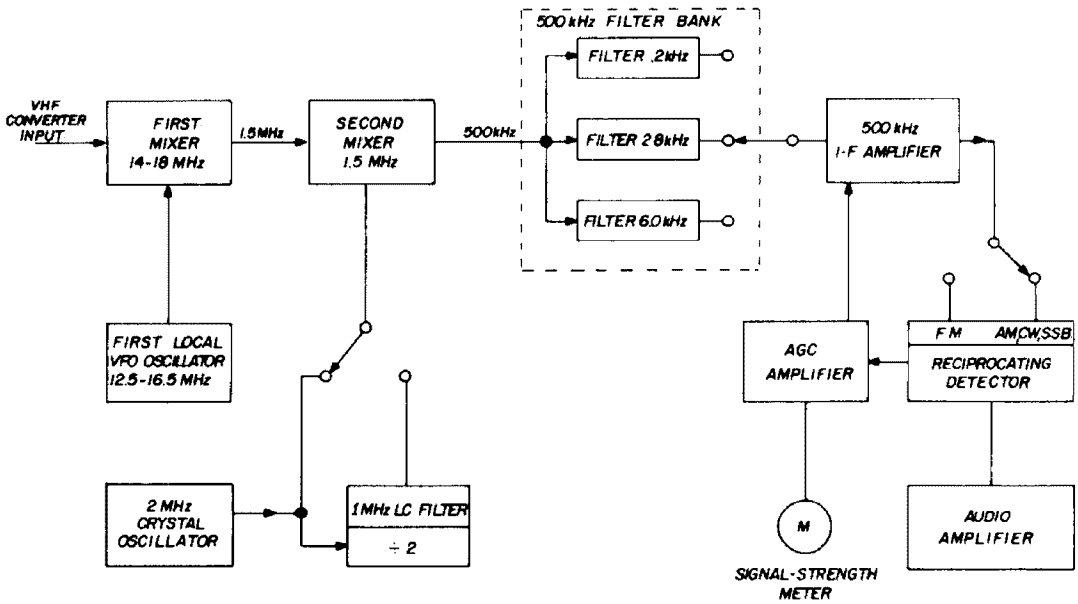


fig. 1. Receiver block diagram. Set is designed for use with 2- and 6-meter converters whose outputs are between 14-18 MHz.

construction

This article was prepared with the serious builder in mind. I've tried to give construction tips and guidelines for those who enjoy constructing radio equipment. You are urged to consult the material listed in the references at the end of the article, which I've chosen to provide further information on working with PC boards and toroid inductors.

The receiver is built on a 6 x 6 x 2½-inch aluminum chassis, which was fitted with a panel and side brackets. The first conversion section (fig. 2) is constructed on a piece of epoxy copper-clad board, which was drilled and fitted with flea clips to support the mosfet RCA 40673 and its input circuits and the coil for the first local oscillator. An MPF102 fet, which serves as the transistor for the tunable oscillator, is also mounted on this board.

When winding the oscillator toroid, first wind the wire on a match stick, which serves as a bobbin and can be passed easily through the core center. Pull the wire as tight as possible. Anchor the wire endings with small pieces of tape, then dope the windings into place.

The tunable oscillator main capacitor is a surplus unit. It was used in a LM or BC221 frequency meter and bears the inscription Cardwell BC11-71-48. It has an excellent loaded gear train and an

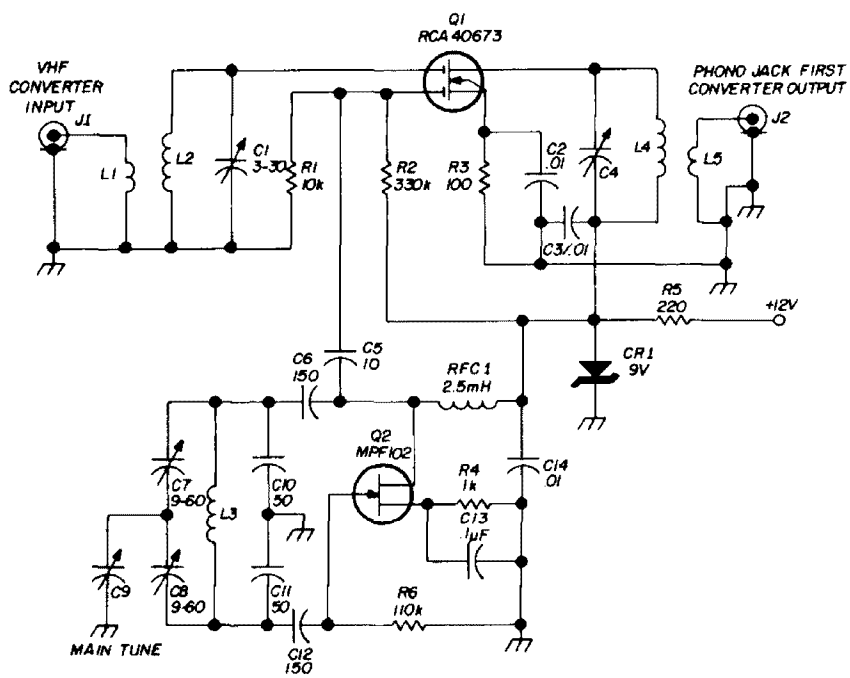


fig. 2. First converter schematic. L1: 2 turns no. 22E. L2: 12 turns no. 22E. Both coils wound on Amidon T-22-2 core. L3: 14 turns no. 26E wound on Amidon T-25-2 core. L4: Miller A123A coil with 6 turns added at bottom end of coil form for link L5.

first converter

The first conversion i-f transformer is located on the same board with the components described above. The input curcuit to the first mixer, which allows either of the converters to be switched in, is a coupling link to allow low input impedances of the vhf converters to match this input. A shaft extends through the front panel so that the input circuit can be peaked across the 14-18 MHz band. The first mixer output coil is link coupled to the second mixer input through a short length of coax.

The possibility of coil interaction is remote, but care should be used in mounting each coil, particularly the oscillator coil, since movement of any parts will cause frequency instability.

extension on its main shaft, which allows a dial-cable pulley to be added. The dial-cable pulley assembly was fashioned on the front panel to accommodate a slide-rule dial with a large calibration area. A piece of graph paper provides the dial division marks, which are calibrated by pencilling in the main divisions.

The oscillator/mixer assembly is mounted in a Zero box.* The box cover is mounted to the main chassis. All component supports and the two tuning capacitors are mounted on stiff brackets; their shafts extend through slots cut into the box. Aluminum deep-drawn boxes are used as shields and compartments throughout the receiver. These boxes

*Zero Manufacturing Company, 288 Main Street, Monson, Massachusetts 01057.

provide rf-tight shielding, which is essential. Sheet-metal screws are used to secure box covers.

If you wish to use a different dial and main tuning capacitor, choose a variable capacitor with double bearings. A Miller 2101 capacitor can be used as a substitute for the unit used in this construction. A dial with a gear reduction may be used in place of the slide-rule dial described here.

second converter

The second converter (fig. 3) is almost a duplicate of the first.* It uses 1.5 MHz as its input frequency. The input to this circuit is fed through a phone jack, which connects to a low-impedance link to the mixer input coil. A short piece of coax connects the first and second converters through this jack.

The second mixer also uses an RCA 40673. The second oscillator is crystal controlled and uses the divider method to generate the local oscillator signal.[2,3] The output of a 2-MHz crystal oscillator is fed into one-half of a 7473 flip-flop, which operates as a frequency divider to provide a 1-MHz signal. The output of the crystal oscillator at 2 MHz, or the 1-MHz output from the divider, is filtered through tuned circuits. Either of the two filtered outputs is presented to the mixer by a selector switch, which allows the lower or upper sideband to appear in the mixer output, which is 500 kHz. If a lower i-f, say 455 kHz is desired, a different crystal oscillator frequency must be chosen, which would be 1955 kHz for the upper sideband and 1045 kHz for the lower sideband. These outputs are the second converter local oscillator frequencies.

The second oscillator and divider are constructed on a piece of copper-clad epoxy board. The crystal oscillator, its tuned output circuit, and the frequency divider are also constructed on a piece of copper-clad epoxy board. Flea clips are used to mount all parts including the 7473 IC.

A Vector pad drill,† used in conjunction with a small drill that is used to cut a pilot hole for the pad drill, is a commercial version of a device described in an earlier *ham radio* article on the construction of instant printed circuits.[4] These tools can be used to cut out copper pads in copper-clad board so that terminals can be fastened to them for easy mounting of components. This technique was employed throughout the entire construction of this project and is highly recommended.

The second LO board is mounted on 4¼-inch standoff bushings within the cover of a 2¼ x 1¼-inch Zero box cover. The cover is mounted on the chassis to the right of the first converter box. Clearance holes through the bottom of the box and the main chassis allow connection to the upper or lower sideband selector switch. The second conversion input transformer, the mixer fet, and the two oscillator filters are in the same shield box. The output of the mixer is fed to a Millen 61455 i-f transformer, which is retuned to 500 kHz by replacing the capacitors presently installed in parallel with the primary and secondary coil with two 100 pF mica capacitors. This transformer, located to one side of the second-conversion mixer shield, provides the signal for the mechanical filters mounted below the chassis directly under the first converter box. This construction allows short leads from the band-width selector switch, SW1, to the filters. A single-stage transistor amplifier is mounted on the back of SW1. This amplifier compensates for filter losses and transforms the filter output impedance to match the two-stage i-f amplifier.

i-f amplifier

The ICs for the i-f amplifier are Motorola 1550Gs. A Millen 61455 i-f transformer, retuned to 500 kHz, is used as an interstage transformer. The output i-f transformer is a toroid. All these components are mounted in a third Zero box in the same manner as the second converter. Amplifier output is by means of a small length of coax to the detector compart-

*A complete parts list is available from *ham radio* for $1.00 and a self-addressed stamped envelope.

†Vector pad cutting tool no. 116.

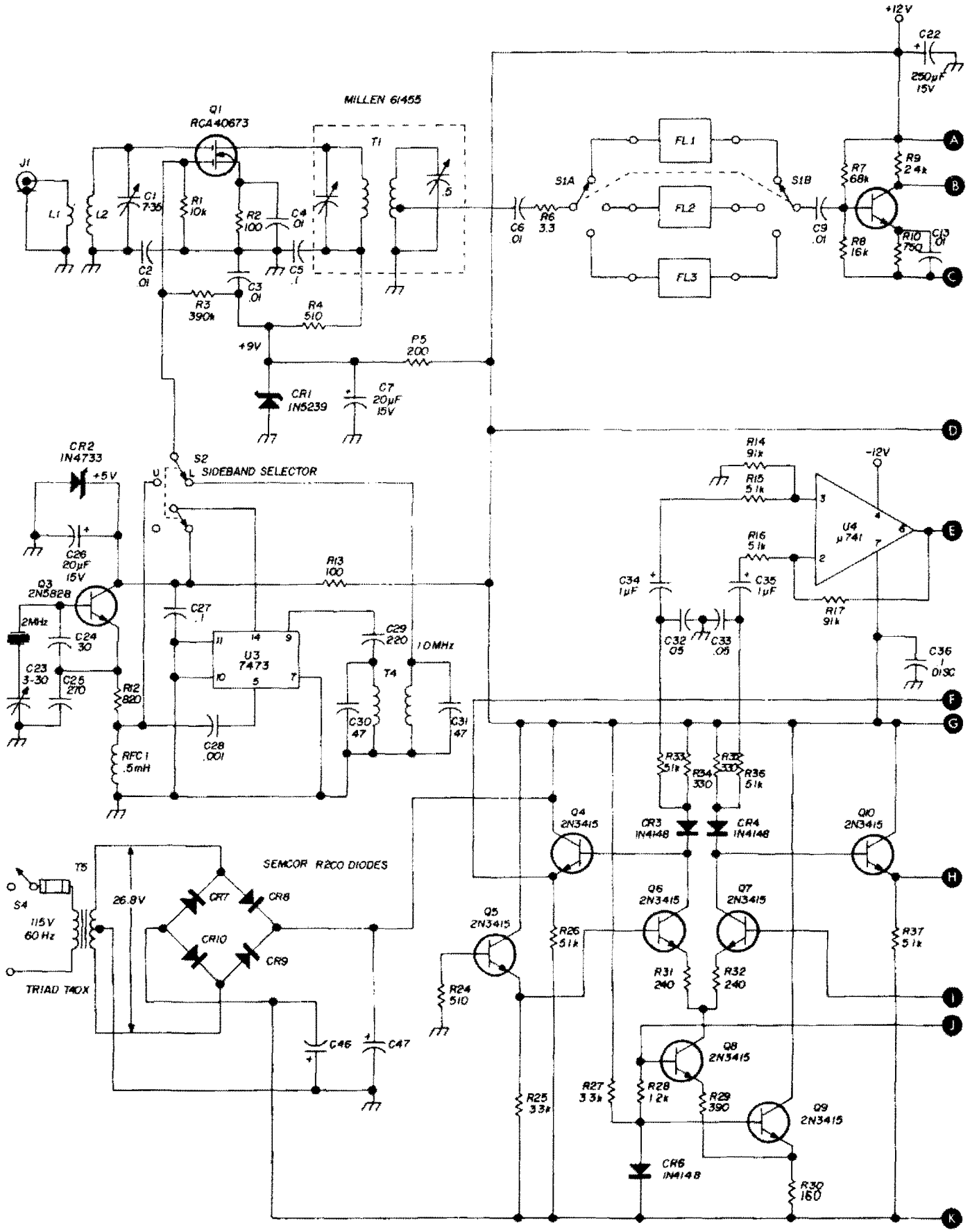
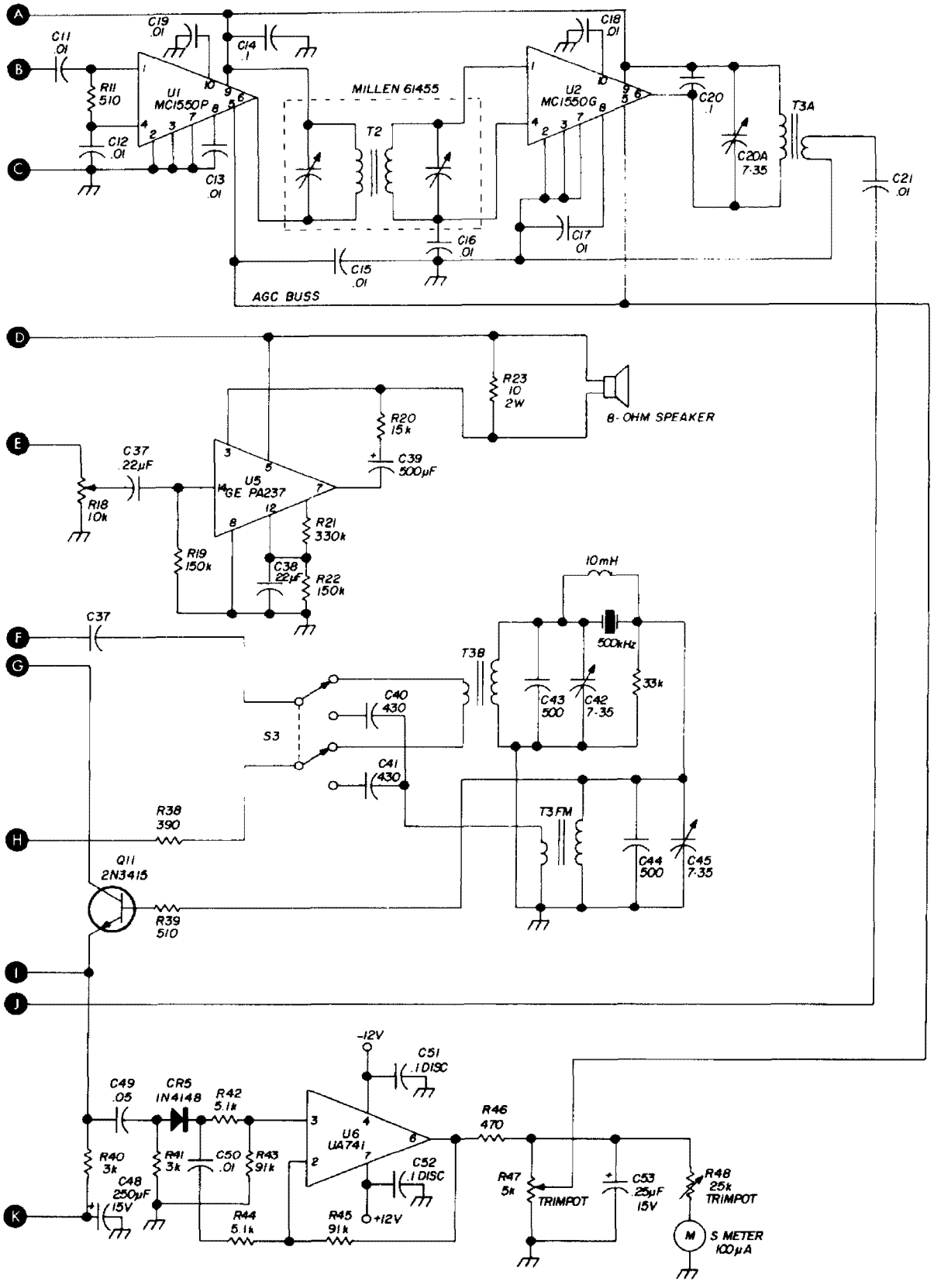


fig. 3. Schematic of 500-kHz i-f strip. L2: 74 turns no. 32E on Amidon T-144-15 core. L1: add 6 turns no. 32E over L2. T3A, T3B, T3FM: primary 14 turns no. 34E; secondary 92 turns no. 34E. All three coils wound on Amidon T-144-15 core. T4: Miller 12W1 i-f transformer.



ment on the bottom left front of the main chassis.

The second detector is a reciprocating detector. This circuit does not require a bfo. It synthesizes a reference signal from the received signal, which serves as a beat-oscillator signal. The reference level is proportional to the average signal received. The circuit does not contain the background hiss prevalent in bfos used with conventional detectors. Further low-noise improvement is due to a nar-

rowband filter employed in the circuit that extracts the reference signal.

A recent investigation on fm, revealed that the reciprocating detector is a satisfactory fm discriminator. As a discriminator it makes its introduction in this unit, which makes possible an all-mode receiving system. By adding a tuned circuit to the components used in the reciprocating detector, it's possible by means of a switch to extract the sum instead of the difference frequency of the output. Suppression of any tendency toward positive feedback and a 90-degree phase shift produces essentially a conventional fm discriminator. In our unit (**fig. 3**) the tuned circuits are designated T3B for a-m, ssb, and CW and T3FM for fm.

All detector components are on a piece of epoxy board, which is mounted on ½-inch bushings fastened to the main chassis next to the mode selector switch. A shield for the detector circuits, made of a 1¼ x 2 x 3-inch box with a removable cover, is mounted over the epoxy board.

Agc voltage is extracted from the reference emitter-follower output in the reciprocating detector, rectified, and applied to an agc amplifier, which assures a wide range of control. An S-meter output is included, but no meter was mounted on the panel for lack of space. The audio amplifier has enough gain to drive a speaker. The power supply shown in the schematic is adequate for the entire receiver.

alignment and test

Alignment procedure is straightforward. A vtvm, rf probe, and signal generator are required.

First determine that wiring is correct and that coil sense is proper. Begin by applying voltage to the first converter. To determine if the first LO is working, place the vtvm rf probe on the drain and rf choke junction. The rf level will be around 3 volts at the low-frequency end of the oscillator range. It will drop off slightly at the high end. Next adjust range-setting capacitors C7, C8 to about 50% closed.

The main tuning capacitor, C9, should be 95% closed. Tune in the oscillator on a receiver or frequency meter; its frequency should be very near 12.5 MHz. If not, carefully adjust the range setters until the signal is audible in the receiver. Now adjust the main tuning dial until the capacitor is about 75% open, where 16.5 MHz will be audible in the receiver or frequency meter.

With dc applied to the second converter and with the vtvm rf probe connected to the arm of the sideband selector switch at the junction of the 15-pF capacitor, determine that the 2-MHz crystal is oscillating by placing the switch in the upper sideband position. Approximately 3 volts will be available here and nearly the same on lower sideband position if T4 is correctly resonated. If not, tune the primary side first, then the secondary for maximum output as indicated on the vtvm.

Switching between U or L should indicate about the same voltage level. These two frequencies will be 2.0 MHz for U and 1 MHz for L. Place the rf probe on terminal 2 of T1, place the sideband selector switch on U, apply a weak 1.5-MHz signal input to J1, and adjust T1 primary for maximum on the voltmeter via the rf probe. Move the probe to the junction of the 3.3k resistor and the arm of SW1A, adjust T1 secondary for maximum and repeak the primary. The transformer coupling should be adjusted to mid position. Now move the rf probe to terminal 6 of U1 and place filter selector switch SW1A, SW1B to no. 3 position, which puts the 6 kHz mechanical filter into the circuit.

Move the probe to pin 6 of U2, adjust the primary coupling of T2 to midway, then adjust the primary and secondary of T2 to midway, then adjust the primary and secondary of T2 for maximum on the vtvm. Adjust C20A of T3A for maximum. Move the probe to the high side of the output link of this transformer, and note that output exists at a 1½ times decrease in level.

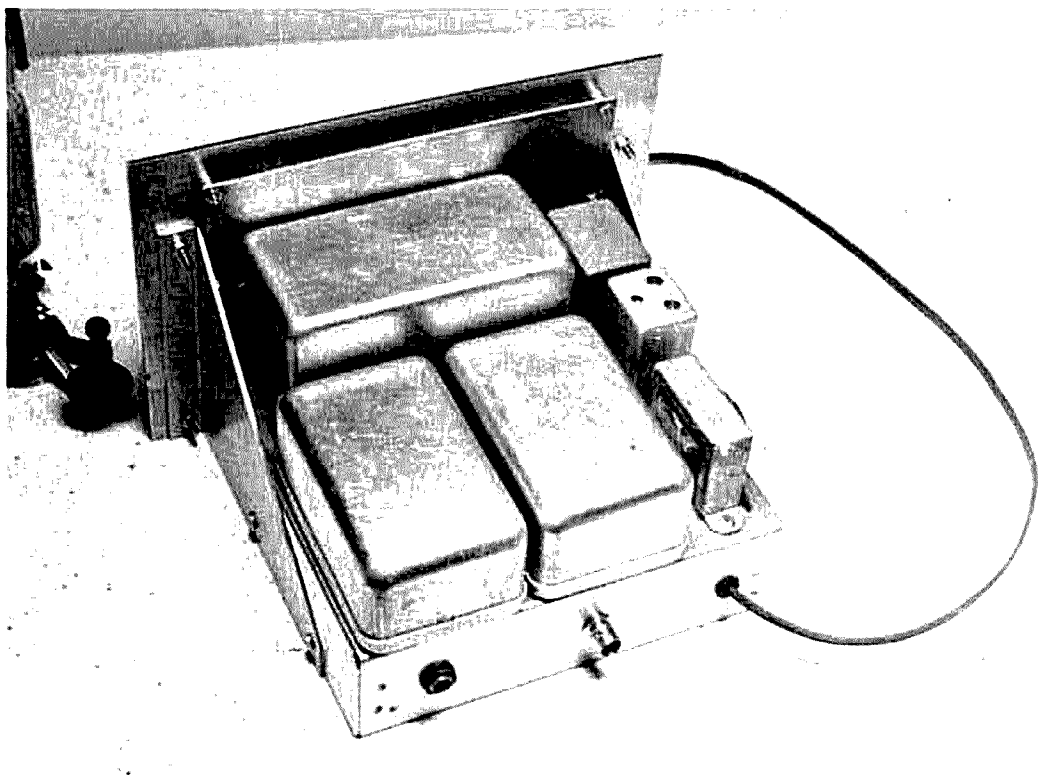
With the rf probe still in the same position, move the bandwidth selector to the 0.2 kHz position and retune each

adjustment for T3A, T2, and T1 for maximum output in that order. At this point less input signal may be required. Decouple the signal generator to a level that ensures limiting has not occurred due to over driving. This signal will be approximately 10 microvolts.

Connect the first converter output to the 1.5-MHz input jack. Tune the main tuning dial to the point determined to be 12.5 MHz when the first converter was aligned. Remove the second converter aligned, and we can proceed to the detector alignment. By now some indication of a signal must be evident from the speaker or phones.

detector alignment

With the sideband selector switch in the L position and the mode selector switch in a-m, CW, and ssb (which is the same switch position), connect a vtvm, set to measure dc at a very low voltage, to the emitter of Q8. Disconnect C21 from



The companion receiver chassis. Battleship construction and shielding make for superior mechanical and electrical stability.

shield and connect a small piece of wire to the first converter input jack. Adjust the second converter input coil tuning capacitor, C3, for an increase in signal as indicated on the vtvm. The rf probe should still be connected to pin 6 of U2. Now adjust the main trim control on the front panel for a further increase in signal. This signal is at 14 MHz. Each megahertz throughout the 14-18 MHz range can be determined by tuning in the beats with the main tuning control. Replace the Zero box cover and the signal should disappear.

The front-end and i-f stages are now the i-f output. A voltage between 100 and 200 millivolts should appear across the R29, R30 combination.

Transistor Q8 functions as a half-wave rectifier as well as a current source; for maximum dynamic range it should draw a small amount of current even in the absence of a signal. The voltage described, therefore, is the result of the current flow across these two resistors. A too-low voltage will cause distortion or even complete silence at low signal levels; conversely, a too-high biasing current will cause a loss of impulse-noise rejection and synchronous bandwidth.

The narrowband filter used in the reciprocating detector is very simple to construct. A 500-kHz crystal is used. Since the bandwidth must be 500 Hz to the 3-dB points, an inductance could not provide sufficiently high Q, so a combination of inductance and the Q of a quartz crystal is used. The crystal is a surplus HC6. The inductance across the crystal tunes out the crystal capacitance so that a uniform band shape is achieved. The input transformer allows the filter to be driven balanced; its unbalanced output is taken from the top of a 33k termination, which drives an emitter follower to the input of the synchronous switch.

filter alignment

To adjust the filter, turn the receiver off as no power is required for this adjustment. Apply a 500-kHz signal to the emitter of Q4, connect a vtvm rf probe to the output of the filter, which should be disconnected from R39, a 510-ohm resistor. Now adjust C42 for maximum signal and tune through the signal several times to determine that resonance has been achieved. Measure the signal generator rf level and compare it with the filter output level; the ratio of the generator output, E_g, divided by the filter output, E_o should be at least 3.5 with the 33k termination in place. The filter bandwidth will be approximately 500 kHz when R38 is 390 ohms.

Reconnect the filter to R39. Reconnect C21, turn on the power, and apply a 14-MHz signal to the converter input jack. A heterodyne will be heard, which will disappear when the main tuning is adjusted through zero beat on this signal. The zero-beat range will have a small area where nothing will be heard; this is the lock in range of the detector. If the beat is not present, reverse the secondary leads of T3A to put the transformer in the correct phase relationship.

final adjustments

To tune up the detector for fm, resonate transformer T3FM to 500 kHz in exactly the same way you adjusted T3A. The exceptions are that the tuning will be a little broad and it is not necessary to plot the E_g/E_o level. The bandwidth will be about 15 kHz.

To adjust the agc and the S-meter amplifier, complete the following procedure: The 14-MHz signal at the converter input must be reduced in level so that it is hardly perceptible. Connect a 0-100 microamp meter to the point marked S-meter in the diagram. The meter should show some indication of noise impulses near its zero point. If such is not the case, adjust R47, a 5k trimpot, until the meter reads zero. Now increase the signal generator output until a 2.8-volt signal is measured at the output of U6, then adjust R48 until the meter is at full-scale deflection.

In my construction this meter was not put on the front panel but is a part of a console, which contains an antenna rotator control. The meter is used to peak signals with a beam, so an external connection is made through a jack at the rear of the receiver.

There's not much more to be said about this receiver except that it fulfills its requirement with vhf converters and will hold its own with my R4A, which I now happily operate on 20 and 15.

I hope this project will be a useful guide in construction if you too become a DX widower.

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ham radio

phase-locked loop AFSK generator

This integrated-circuit phase-locked loop AFSK generator provides excellent long-term frequency stability

Daryl J. Duffin, K7ZOF, 255 Iowa, Ogden, Utah 84404

After trying several circuits for a stable AFSK tone generator, and meeting with various forms of failure, I finally decided to do what I should have done in the first place — use a phase-locked loop function generator, the Signetics 566. This little device puts out both triangle and square waves up to about 1 MHz. The frequency of the 566 is programmable by a resistor (R8), capacitor (C1), and voltage or current at pin 5.

In this application, the AFSK frequency is set to 2125 Hz *(mark)* by R1-R5 and R10. Then the voltage at the modulation input is changed sufficiently to move the frequency up to *space* (2295 or 2975 Hz), or to 2225 Hz for narrow-shift CW identification. This is accomplished by feeding the FSK keying voltage from the RTTY terminal unit to a transistor inverter stage which keys the phase-locked loop.

The keying transistor, Q1, is cut off in *mark,* allowing R1-R5 and R10 to set the frequency. In *space,* the keying transistor is biased on, pulling current through either R2-R6 or R3, lowering the voltage at pin 5; this raises the output frequency to *space.* If the key is closed, the frequency is similarly raised through R4 and R7.

Since the ST-5 and ST-6 both have plus and minus power supplies, and the 566 IC is designed to operate that way, the pair are a natural for each other. Although the 566 will operate with up to

24 volts, this is its maximum rating, so 4.7-volt zeners were used to drop the voltage to the device. The circuit will also work with a single +12 volt supply by grounding the minus terminal, feeding +12 volts to the positive supply, and juggling the frequency-setting resistors.

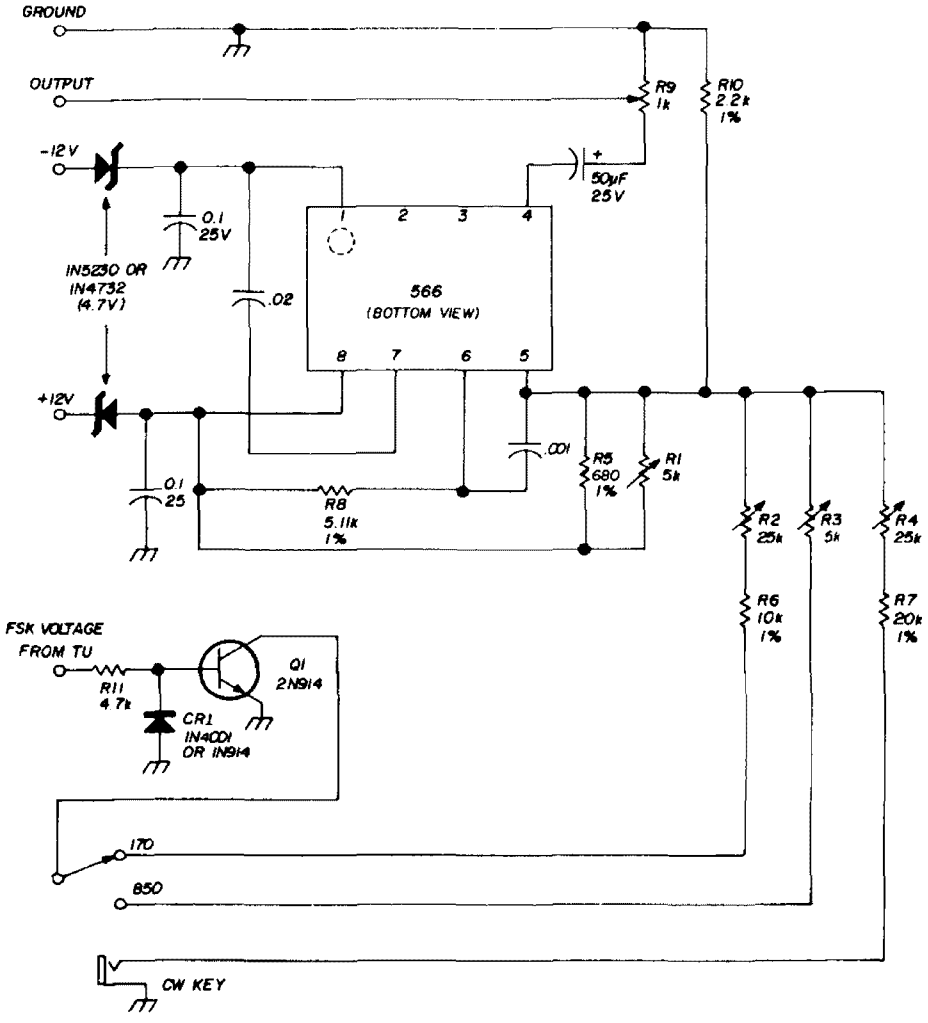


fig. 1. Circuit for the phase-locked loop AFSK generator. The IC is a Signetics 566. An alternate keying circuit is shown in fig. 2.

No problems were experienced with the triangular output voltage, since the bandpass circuits of any rig used with this AFSK generator will remove the high-frequency component of the oscillator.

alignment

To set up the generator, use a frequency counter or well-aligned terminal unit, and follow the following steps:

1. With the terminal unit in *mark*, or a negative or zero voltage at the keying input and CW key open, adjust R1 for 2125 Hz at the output.

2. With CW key closed, adjust R4 to provide 2225 Hz at the output.

3. With the terminal unit in *space*, or +10 volts or so at the keying input, and the mode switch in the 170 shift position, adjust R2 for 2295 Hz at the output.

4. With the mode switch at 850 shift, adjust R3 for 2975 Hz at the output.

5. With the rig that will be used, adjust R9 to the proper operating level.

You will note the odd resistor values in the circuit; I used surplus precision resistors for thermal stability. However, carbon resistors would probably suffice. I used 10-turn wirewound pots for the

adjustable resistors (except level). It might pay to experiment with fixed resistors in parallel with the pots to narrow the adjustment range and alleviate the problem of the pot slider hitting two wires, each of which may be to either side of the desired frequency.

This AFSK generator was designed to use the FSK voltage output of the ST-5 and ST-6. This is -10 volts on *mark*, and +10 volts on *space*. I initially used the alternate keying circuit (fig. 2) with the RTTY keyboard itself keying the transistor, but this system didn't work too well due to the unstable keyboard resistance and the requirement for two separate loop circuits.

I built the generator on perfboard the same size as the ST-6 boards, using copper-foil tape. The layout is not critical, but mount the pots so they can be easily adjusted. My copper-foil layout is very similar to the schematic diagram.

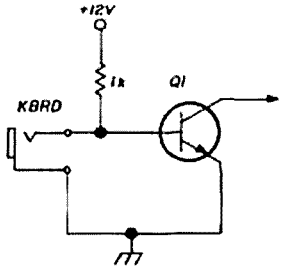


fig. 2. Alternate keying circuit for the phase-locked loop AFSK generator.

I have used this circuit for over a year and have not had to readjust it after it was set initially. I use this circuit on both vhf fm as AFSK, and on low bands by just feeding the output signal into the microphone jack.

My thanks to Al Crapo for doing the complex math needed to come up with the resistance values and circuitry required. Without that, I would still be diddling with resistor values!

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ham radio

Reception

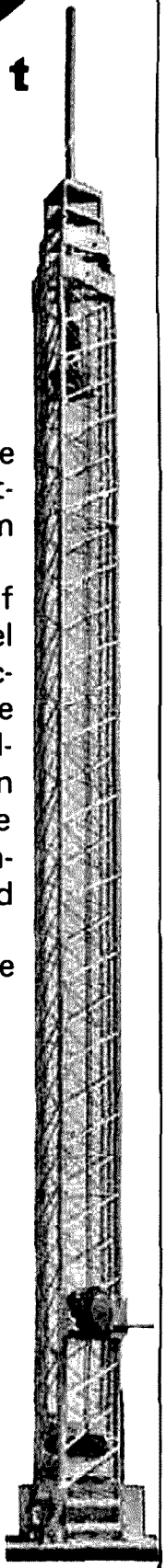
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proper LC ratio in the amplifier tank, may occur. In such cases, an antenna that attenuates harmonics is highly desirable. Usually, this will be a sharply tuned resonant antenna (fig. 2), and the addition of an antenna tuner will help. If the system is matched properly, a low-pass filter may help.

antenna installations

Coax cable can radiate like a bearcat. If the antenna mast is placed at the side of the house, is hollow, grounded, and the coaxial cable is run inside it, radiation will be reduced greatly. Radiation is then in the horizontal plane, assuming a dipole or beam is used. If a balun is used at the antenna feed point, a better balance and less trouble may be expected. Such an installation may give as much as 30 dB discrimination when referred to the vertical downlead of a TV receiver antenna installation using 300-ohm twin lead. A trap at the TV receiver or a high-pass filter at the TV set will help (fig. 2). It helps public relations if you pay for it, but let a TV serviceman install it. Otherwise, if anything goes wrong with the TV set, you will be the culprit and will be expected to fix it or foot the bill.

The troubles are usually bad on 6 and 10 meters, and sometimes on 15. It can

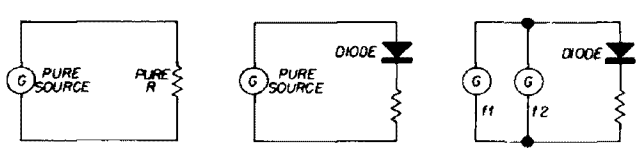


fig. 1. A signal source connected to a pure resistive load produces no harmonics. A diode connected in series with the load produces harmonics and harmonics plus cross modulation.

happen on any band, but I think sometimes the reason amateurs use 75 meters so much is because of the relative freedom from RFI on this band. With a-m, you are easily identified. With ssb or CW identification is more difficult, but a mast in your backyard is a dead giveaway. An inconspicuous antenna in crowded communities is highly desirable.

Running coaxial cable in the ground will help reduce stray radiation. The antenna should be sharply tuned and resonant at a single frequency rather than a multi-band type. With the coax shield

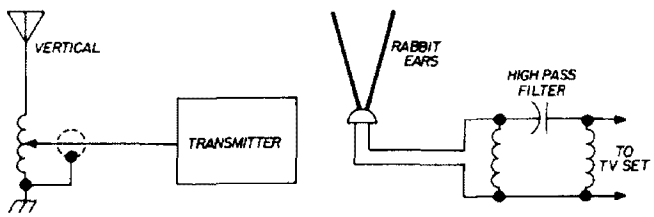


fig. 2. Selective circuits in the transmitter output reduce harmonics. A high-pass filter between a TV set and the tuner input is effective in attenuating strong amateur signals at the transmitter fundamental frequency.

grounded, the cable tends to act like a low-pass filter. Running the cable in a piece of galvanized pipe will also help reduce RFI (fig. 3).

power-circuit coupling

Inside the home, coupling between ordinary lampcords and power wiring

Editor's note:

Despite the vast improvement in electronic communications equipment design over the past few years, the problem of amateur transmitter interference with home-entertainment devices is still much in evidence. Thanks to the efforts of industry in this country and concerned amateur groups, TVI isn't nearly as serious as it was 20-25 years ago. Today, TVI has been replaced by a bugaboo known as TXI, which includes interference from ham transmitters with equipment such as f-m broadcast receivers, stereo record players, and even hearing aids. Amateur transmitter interference with public telephone equipment is very much a problem. These interference modes may be lumped under an all-inclusive category known as RFI — radio-frequency interference. This article presents some suggestions for handling the problem. WA3NFU doesn't pretend to provide solutions for every type of RFI. Rather, a compendium of basic RFI causes and cures is given; and the knowledgeable amateur, armed with this information, should be able to resolve his particular RFI problem.

should be minimized. Placing the transmitter near a window, and having a short direct run for the antenna cable to the outside of the building, will tend to minimize stray coupling to power circuits. Of course, if the wiring of the outside power system is open, on poles, and not buried, and you radiate toward it from the antenna, the rf will feed right

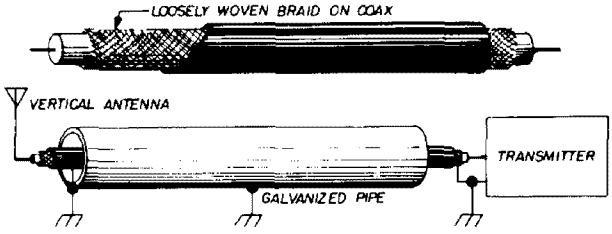


fig. 3. The braid in most coax cable does a poor job of RFI shielding. Burying the coax in the ground or installing it in a pipe provides effective shielding for RFI.

back into the house and may also get into telephone circuits. Installing the antenna on a high mast and using horizontally polarized radiation may help. Since the power wires are horizontal and may run for miles, a vertical antenna may actually be a better RFI solution because of reduced coupling. The base of the vertical can be at ground level, making adjustments and tuning more convenient. Each case is unique and experiments are necessary to find the best solution. Using a vertical ground plane on a mast is likely to be the worst case.

A neglected part of the transmitter installation is the power cord from the transmitter to the electric outlet. Preferably, this cord should have an rf filter and the wiring should be shielded and grounded. A ground may be made to the BX cable in the house wiring and also to a ground rod. A heavy, low-resistance conductor should be used. If the house wiring is old, connections and joints should be examined for corrosion. Corroded joints form diode rectifiers, and you know what that can cause, especially with strong rf currents.

If the transmitter runs high power, switching it on may cause the lamps in lighting fixtures to dim because of poor line-voltage regulation. If this problem occurs with low or medium power, rf may be in the power circuit. If the lamps, especially fluorescents, light without being switched on you'd better check for rf in the power system. If a neon lamp or fluorescent lamp glows when placed near an rf line, the presence of rf and standing waves on the line is assured. Often this means RFI.

transmitter problems

Let's now examine what is probably the most predominant cause of RFI — the transmitter. The sketches in figs. **4** and **5** illustrate some of the more obvious problems, which are discussed below.

Some amateurs have a habit of not using all the screws when reinstalling a bottom plate or cover of a transmitter. This may reduce shielding effectiveness and cause stray radiation. The screws should be in and reasonably tight. A ground conductor should be run from the ground connection of the transmitter to a

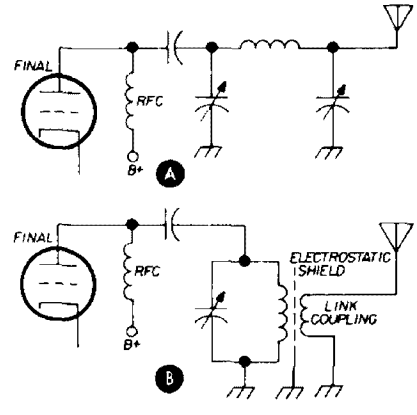


fig. 4. Typical pi network is effective in attenuating harmonic radiation only if reactance is low or zero. An electrostatic shield is useful with link-coupled output circuits but presents mechanical problems.

solid ground rod buried at least six feet deep, preferably in moist soil. This is important from a safety standpoint as well as for minimizing RFI.

If the antenna and ground system are all right and RFI troubles still persist, the fault may be due to a misadjustment of the transmitter or a defect in it. Over-

modulation, for example, can cause a host of troubles. Modulation can be checked on a scope or a simple carrier-shift indicator. If the final amplifier is a class-C stage and is not neutralized properly, RFI may result. How many amateurs check the neutralization? Techniques are covered in the ARRL Handbook and elsewhere. Usually this is the last thing to be done and might well be the first.

If the drive for the final stage is marginal due to poor transmitter design, misadjustment, or a fault in a preceding stage, the final may be struggling so hard that its output waveform is highly distorted. A class-C stage by its very nature is a harmonic generator. Many transmitters use a single-ended final, whereas a push-pull final would help to reduce harmonic output. With a single-ended stage it's especially important that the final be tuned properly. An antenna tuner is a definite advantage since it increases the output circuit selectivity.

Proper LC ratio is also important, not

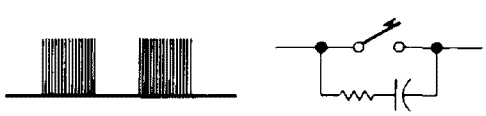


fig. 5. Pulses from CW modulation may be shaped by a simple RC filter.

only from the standpoint of tube efficiency, but from the standpoint of reducing harmonic output. With the transmitter output fed to a shielded dummy load, harmonic output can be checked on a receiver or other suitable device.

If you buy the transmitter or transceiver, you're stuck with the original design. However, if you build your own, you can design circuits that will minimize RFI. All the design data is in the ARRL Handbook and numerous other standard texts. At one time, for example, link coupling between rf circuits was widely used. Now the final amplifier is coupled to the antenna circuit through a pi network. This system is simple but not too good from an RFI standpoint. An electrostatic shield placed between primary and secondary circuits eliminates capacitive coupling, but it is difficult to implement.

keying

When CW is used, the tendency to generate RFI is even worse than with a-m

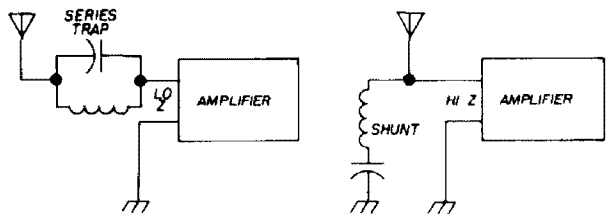


fig. 6. Application of filters for use with low- and high-impedance-input amplifiers.

or ssb. With CW you have a step-function signal, or transient, similar to a radar pulse. To minimize such interference, key-click filters are highly desirable as well as push-pull final stages.

Grid-block keying uses the keyed tube as a switch. Using a shielded keying relay is effective in RFI reduction. If a low-power keyed oscillator is used, followed by several stages of rf amplification, RFI will be less than if a high-power final is keyed. With the key up, in any case, there should be negligible radiated rf.

filters

In general, a series LC filter should be used where the load impedance is low, such as with a 50-75 ohm receiver input. The high-Z circuit should be used in series with low Z (fig. 6). If the receiver input Z is high, the shunt across it should be low Z. Passive filters may give poor attenuation because of mismatching. A shunt high-Z filter may consist of a small capacitor that bypasses rf at the input to a high-gain audio amplifier. Such a filter will eliminate rf rectification or greatly reduce it. A quarter-wave transmission line, which acts as a short circuit across the input of an rf or af amplifier, is often used.

ham radio

how to use ferrite beads

How to choose ferrite beads so they do the job you intend them to

M. A. Ellis, K1ORV, 61 Marlboro Road, Sudbury, Massachusetts 01776

This is dedicated to those home builders who may have gotten into trouble while using ferrite beads in an attempt to stabilize a troublesome circuit. Many times the problem was not resolved, and occasionally it even got worse when the bead was installed. This has led to a lot of head scratching by the hams involved.

Ferrite beads can be a great aid when they are understood and properly used. This seems to be the problem. Most of us merrily install them in the circuit without being certain of their effect or of what we really expect the bead to do for us. Thus, when the desired signal is greatly attenuated or the undesired one not nearly enough, we try another type bead, more beads (or less), until the circuit seems to be working right. If we can't make the circuit work, we remove the beads and try other measures; or we live with the original problem.

What has happened? Did we use a bead with too much or too little attenuation, or one having incorrect characteristics for our circuit? All of these must be considered if we expect equipment performance to match our expectations.

bead characteristics

A ferrite bead is not a simple device but a rather complex one consisting of both resistive and reactive elements. In fact, the simple equivalent circuit of the bead shows a resistor in series with an inductance as shown in **fig. 1**. The impedance of the bead at any frequency is found by solving the equation

$$Z_b = \sqrt{R^2 + X^2} \qquad (1)$$

Since you are dealing with reactive devices, they must be handled with care – otherwise you may get into real trouble.

Can a logical plan of action be established to determine how and where to use beads? I think so. Let's take a typical circuit problem and develop a method for selecting and using the proper bead to do the job. The circuit is shown in **fig. 2A.**

The source impedance, Z_s, is 50 ohms

fig. 1. Equivalent circuit of a ferrite bead includes both series resistance and series inductance.

*-6 dB = 20 $\log_{10}$ (50 + 50)/(50 + 50 + Z_b)
Dividing by 20 gives
-0.3 = $\log_{10}$ (100)/(100 + Z_b)
Taking the antilog of -0.3, we have
0.5012 = 100/100 + Z_b, therefore
50.12 + 0.5012Z_b = 100 so
Z_b = (100 - 50.12)/0.5012 =
49.88/0.5012 = 100 ohms

and the load impedance, Z_L, is also 50 ohms. You are experiencing a parasitic oscillation at 100 MHz that is reaching the load. However, it can do no harm if it is reduced by one-half (6 dB). Therefore, you need to add a bead to the circuit that will reduce the undesired signal by this amount. The new circuit is shown in **fig. 2B.**

choosing a ferrite bead

With a ferrite bead in the circuit, circuit losses are increased. Cowdell has shown how to determine this insertion loss by using the ratio of load voltage with (vo) and without (v1) the new impedance.[1]

$$\text{Insertion Loss Ratio (ILR)} = (E/vo)/(E/vl) = vo/vl = (Z_s + Z_L)/(Z_s + Z_L + Z_b) \quad (2)$$

$$\text{ILR (dB)} = 20 \log_{10} (Z_s + Z_L)/Z_s + Z_L + Z_b) \quad (3)$$

In fig. 2 Z_s = 50 ohms, Z_L = 50 ohms and the desired loss ratio is 6 dB. You are looking for Z_b, the bead impedance to add to the circuit to attenuate the 100 MHz signal by 6 dB.

Solving for Z_b gives -6 db = 20 $\log_{10}$ (50 + 50)/(50 + 50 + Z_b) = 100 ohms.*

The task now is to select one or more beads having a total impedance at 100 MHz of approximately 100 ohms.

If the bead characteristics are given in terms of impedance versus frequency, there is no problem. Look at the graph in **fig. 3** and see that four beads will present 100 ohms impedance at 100 MHz. Keep in mind that the desired signal will also be reduced somewhat unless it is at a very low frequency.

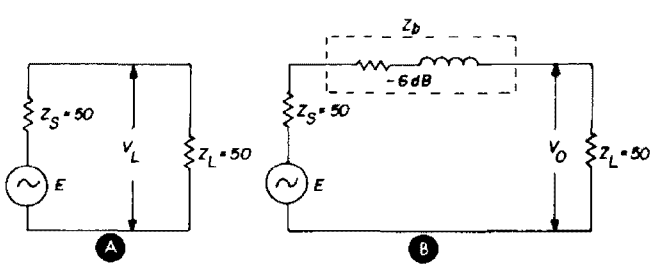


fig. 2. Typical circuit impedances without ferrite beads is shown in (A). When a ferrite bead is used (B), the normal circuit impedances, Z_S and Z_L, are increased by the impedances of the bead, Z_b.

For example, if the circuit is working at 7 MHz, the graph indicates that one bead has an impedance of 9.5 ohms at that frequency. Hence, the total impedance of four beads would equal 38 ohms. The object is to find a bead (or combination of beads) that will give the

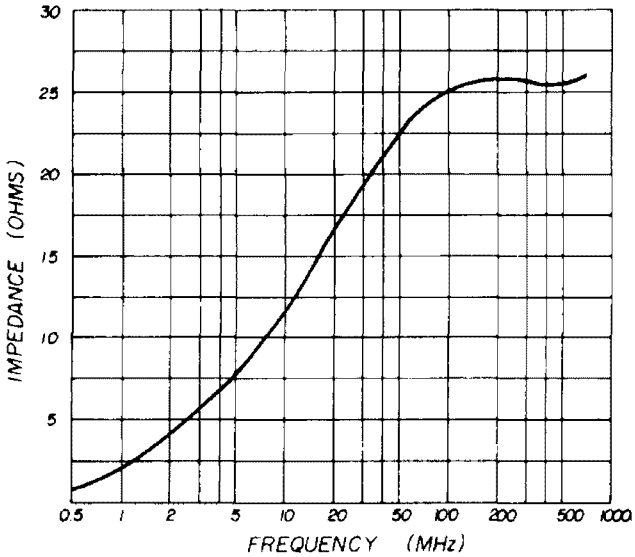


fig. 3. Characteristics of a Ceramag 7D ferrite bead plotted in terms of impedance vs frequency.

desired attenuation of the spurious signal while at the same time giving minimum attenuation to the desired signal.

If the bead characteristics are given in terms of resistance and inductance then you must resort to some elementary mathematics to convert these to impedance values. Looking at **fig. 4** you can see that the resistance is 24 ohms at 100 MHz and the inductance is 0.01 μH.

First, solve for X_L. This can be done by using a reactance chart[2] or you may elect to work the problem mathematically. In any case, X_L = 6.28 ohms; therefore, from **eq. 1**, Z = 25 ohms.

Fig. 5 gives the values of resistance and inductance for Ceramag 7D material* of a certain bead size. Impedance vs frequency for the same bead is shown in the graph of **fig. 3.**

For those of you who have come this far, the graph in **fig. 4** based on **eq. 3** may

*Ceramag Engineering Department, Stackpole Carbon Company, St. Marys, Pennsylvania 15857.

be useful. It allows you to quickly read the impedance required to get bead insertion losses from 1 to 40 dB.

In this graph the source impedance plus load impedance ($Z_S + Z_L$) was assumed to be 50 ohms. If $Z_S + Z_L$ equals another value, the bead impedance, Z_b, read along the horizontal axis may correctly be changed the same amount.

For example, if you want the impedance required for an insertion loss of 6 dB in a circuit having $Z_S + Z_L$ equal to 100 ohms, double the figure of Z_b at the 6 dB point (50 ohms to 100 ohms). This is because the value used for $Z_S + Z_L$ in this example (100 ohms) is twice that used for calculating the curve on the chart (50 ohms). If $Z_S + Z_L$ equals 200 ohms, multiply the Z_b value by 4.

Suppose you have a circuit with $Z_S + Z_L$ = 100 ohms and want bead attenuation in the circuit of 10 dB. From fig. 4 Z_b = 108 ohms. Double that to 216 ohms which is the value to use for 10-dB signal reduction. Now you can select a bead or combination of beads that will provide 10 dB of rejection at the frequency of interest.

summary

To summarize, you must define the problem before deciding what measures to take. First of all, do you have a problem that a bead can solve? If the answer is yes, then how much attenuation is required, and at what frequencies?

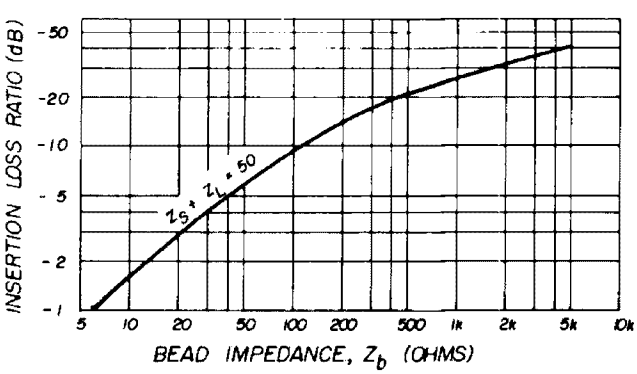


fig. 4. Graph of insertion loss ratio (ILR) vs bead impedance for 50-ohm systems ($Z_S + Z_L$ = 50 ohms). Graph may be used for other system impedance values by use of a simple factor, see text.

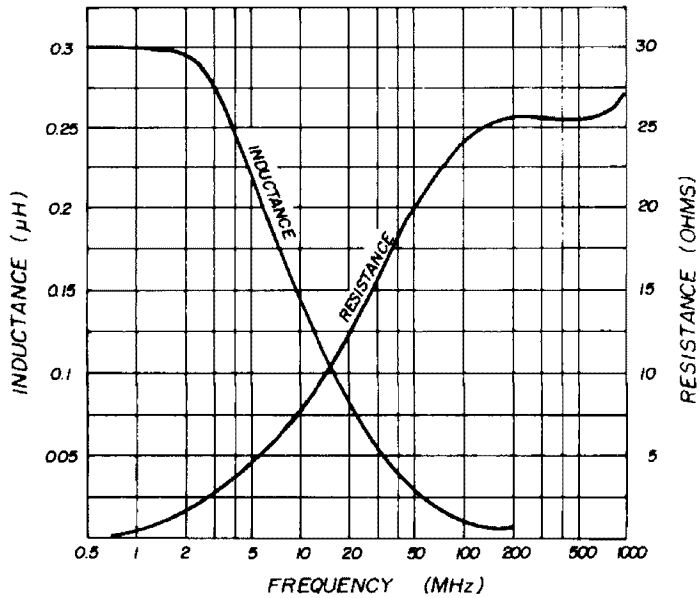


fig. 5. Bead characteristics plotted in terms of resistance and inductance vs frequency. This type of graph is very useful when choosing ferrite beads for your own circuits.

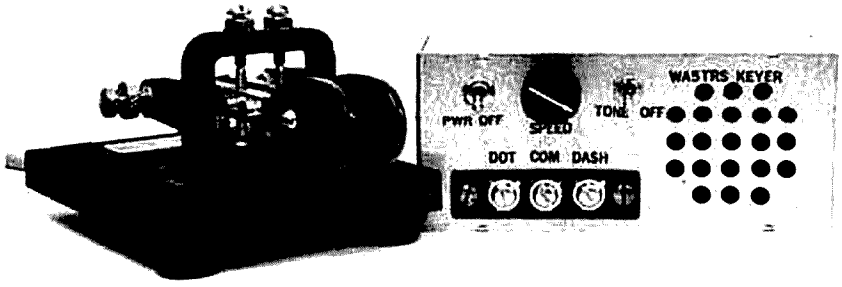
Bead characteristics are then reviewed to determine which ones will give the desired results. It is generally wise to use the smallest size and the fewest in number that will get the job done.

For those readers who may be interested in pursuing the subject of ferrites and ferrite beads, references 1 and 3 through 6 are suggested for additional reading.

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ham radio



simple integrated-circuit electronic keyers

Two simple electronic keyer circuits based on the new Signetic NE555 timer IC

Joe Buswell, WA5TRS, Post Office Box 10674, Midwest City, Oklahoma

For some time I have been intending to learn how digital integrated circuits are used and what their characteristics are. Rereading a series of articles in *ham radio*[1] aroused my curiosity and the discovery of a new device pushed me off dead center and got me started. The new device is the Signetics NE555 timer integrated circuit. The NE555 can be used in numerous applications requiring high stability over time periods from microseconds to one hour.

The most interesting feature about the device is its price, currently about a buck. The NE555 consists of a flip-flop controlled by two comparators. The flip-flop drives two outputs. One is the output which is used to control external circuits; the other is used to control charging and discharging current on a timing capacitor. Fig. 1 shows the basic pin connections of the NE555 and external components used for astable multivibrator operation.

operation

Free-running oscillation of the NE555 occurs when pins 2 and 6 are connected and pin 4 voltage is higher than 0.7 V. If the voltage on the capacitor is less than 1/3 Vcc, the comparator connected to the trigger toggles the flip-flop, causing the output, pin 3, to go to logic high, and causing discharge, pin 7, to go to a virtually open circuit. The capacitor now charges from Vcc through Ra and Rb. When the threshold voltage, pin 6, reaches 2/3 Vcc, the second comparator toggles the flip-flop causing the output to go to logic low and discharge to go to

ground. The capacitor now discharges through Rb to pin 7 until the voltage at pin 2 drops to 1/3 Vcc. At that time, the flip-flop retoggles and a new cycle begins: From the Signetics data sheet,[2] the time the output is high is described by

$$T_{high} = 0.685(Ra + Rb)C \text{ (seconds)}$$

The time the output is low is

$$T_{low} = 0.685(RB)C \text{ (seconds)}$$

The total time period is

$$T_{high} + T_{low} = 0.685(Ra + 2Rb)C \text{ (seconds)}$$

morse code generation

If the circuit of **fig. 1** is constructed with Ra small compared to Rb, the output will be a square wave of 50% duty cycle. This is the requirement for dots. The only problem with keying the circuit up from quiescent with the available control ports is that the voltage on pin 6 must rise from nearly zero to 2/3 Vcc on the first dot and rise from 1/3 Vcc to 2/3 Vcc on the succeeding dots. Therefore, the first dot is 40 or 50 percent longer than its successors.

The bias arrangement of **fig. 2** is used to hold the voltage on pin 6 slightly higher than 1/3 Vcc during rest, allowing it to fall below the required threshold during key down and form an almost perfect first dot.

The control, R2, is used to adjust the threshold voltage slightly higher than the internal threshold voltage and allow for the voltage drop of the diode. R2 should

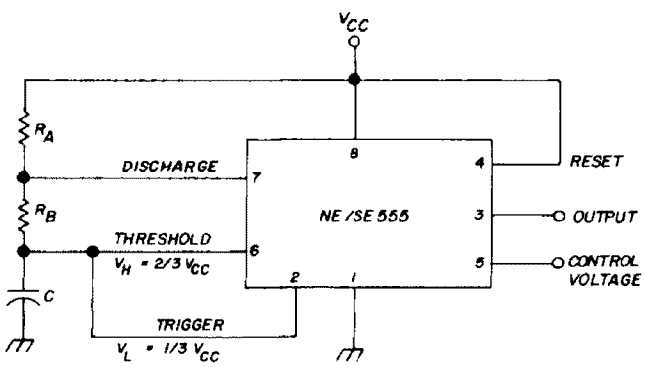


fig. 1. Astable multivibrator circuit using a Signetics NE555 IC.

be approximately equal to the value of R1 to allow sufficient adjustment range. R1 should be included to limit the current drain on the power supply when the key is closed and R2 is inadvertently adjusted to maximum voltage. Diode CR1 is included so that the bias circuitry

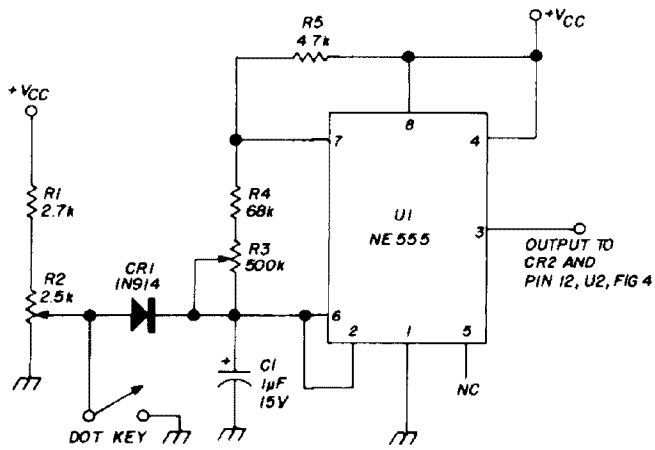


fig. 2. Dot generator using the NE555 IC. Potentiometers R2 and R3 should have linear taper (see text).

won't disturb operation of the timer during the charge-discharge cycle.

dash-dot ratio

The generation of dashes can be accomplished by various methods, of which two are described here. The first, which is generally used with digital logic circuits, is to frequency divide the dot pulse train by two and add the results to the dot pulse train as shown in **fig. 3.** The length of the dash formed by this technique is always equal to three times the length of one dot. The method also provides a dash space equal to one dot-length, which is correct.

This method is used in the solid-state keyer described in the ARRL Radio Amateur Handbook using RTL devices. With a little juggling of components, the more modern and less power-consuming TTL devices can be used.[3] The circuit of **fig. 4** results from a desire to minimize parts. The JK flip-flop, a 7470 IC, was selected to perform the dividing because it has a built-in inverting amplifier on one of the J inputs (as well as on one of the K inputs) which allows grounding of, rather

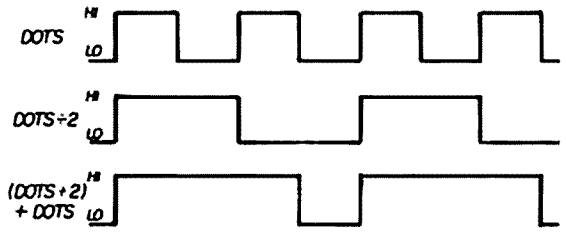


fig. 3. Timing diagram showing formation of dash length that is three times the length of one dot.

than supplying logic high to, the J input to activate the divider. Adding the flip-flop output and dot output is accomplished by diodes CR2 and CR3.

Closing the dash key grounds the timer control through CR4 and initiates a dot. At the same time the timer is initiated, the $\overline{J}$ is brought to logic low which readies Q to go high on the rise of the clock pulse. It is important that $\overline{J}$ be low *before* the clock pulse goes high. This is normally accomplished by the slight time lag caused by discharge of the timer capacitor, C1, fig. 2, down to the trigger voltage where the output toggles high.

If the bias voltage set by R2 is too close to the trigger voltage, there may be insufficient time between grounding of $\overline{J}$ and the arrival of a clock pulse. The result is a dot followed by a dash. The remedy is to increase bias voltage by adjustment of R2 until reliable dash operation is accomplished without excessive delay in initiation of characters after key closure.

After the dash starts the key can be opened because $\overline{Q}$, in the low state, will keep the timer running through the second dot initiation by virtue of CR5. At the end of the first dot the Q output of U2 is left high. The leading edge of the second dot resets Q low but the dash continues for the duration of the dot. The timer cannot start another dot (or dash) until the time lapse of a space has occurred. Also, a dot cannot be changed into a dash while the clock pulse on U2 is high. Therefore, the keyer is self completing.

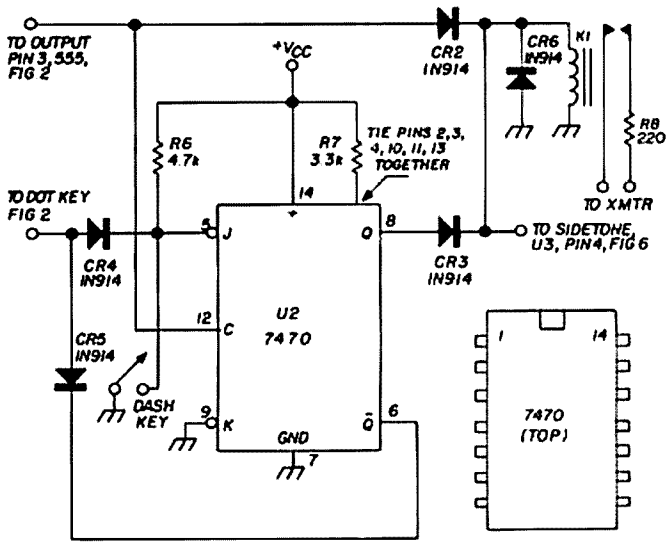
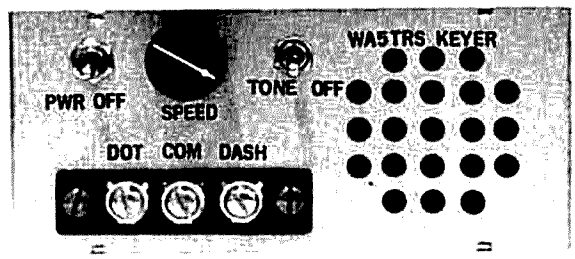


fig. 4. Dash generator circuit. Relay K1 is a 1A5AH, manufactured by the Electronics Application Company, 2213 Edwards Avenue, S. Elmonte, California 91733.



Simple electronic keyer has speed range from 4 to 26 words per minute and has built-in sidetone monitor.

keying speed control

The speed range of the keyer with the components shown is four to twenty-five words per minute. It may be perferable, to suit the builder, to use values of 250k and 33k for R3 and R4 respectively (fig. 2). This will give a speed range of 6 to 50 wpm. Use a reverse log taper control for R3 if it is available. If not, an audio taper control can be wired backwards (CCW rotation increases speed) to help linearize speed vs shaft rotation.

variable dot-dash length

A second keyer circuit which was built to explore the use of logic control circuits and exploit the capabilities of the NE555 timer is shown in fig. 5. In this circuit, when no character is being formed both inputs of gate A are high, forcing the output low. Gate B, wired as an inverter,

has a high output. R1, acting as a voltage divider, places a bias on pin 6 of U2 which prevents the internal flip-flop from toggling, leaving it in its last state which is low on output, pin 3. The low imposed on one of the inputs of gate C forces it high. This high is placed on one of the inputs of gate D. The other input of gate D is also high so the output is low. This latches gate C output high.

When the dot key is closed the output of U2, if it is low, is impressed on one

voltage. Since the output of gate C is high, there is very little voltage drop through R2, so pin 7 assumes nearly Vcc potential. The rate of charge is thus dependent on the values of R4 and R3. When the voltage on pin 6 reaches 2/3 Vcc, the flip-flop toggles, causing pin 7 to go to ground and pin 3 to go to logic low. The current through R2 is increased because of grounded pin 7. C2 now discharges through R3 and R4.

If the dot key is still closed, gates A

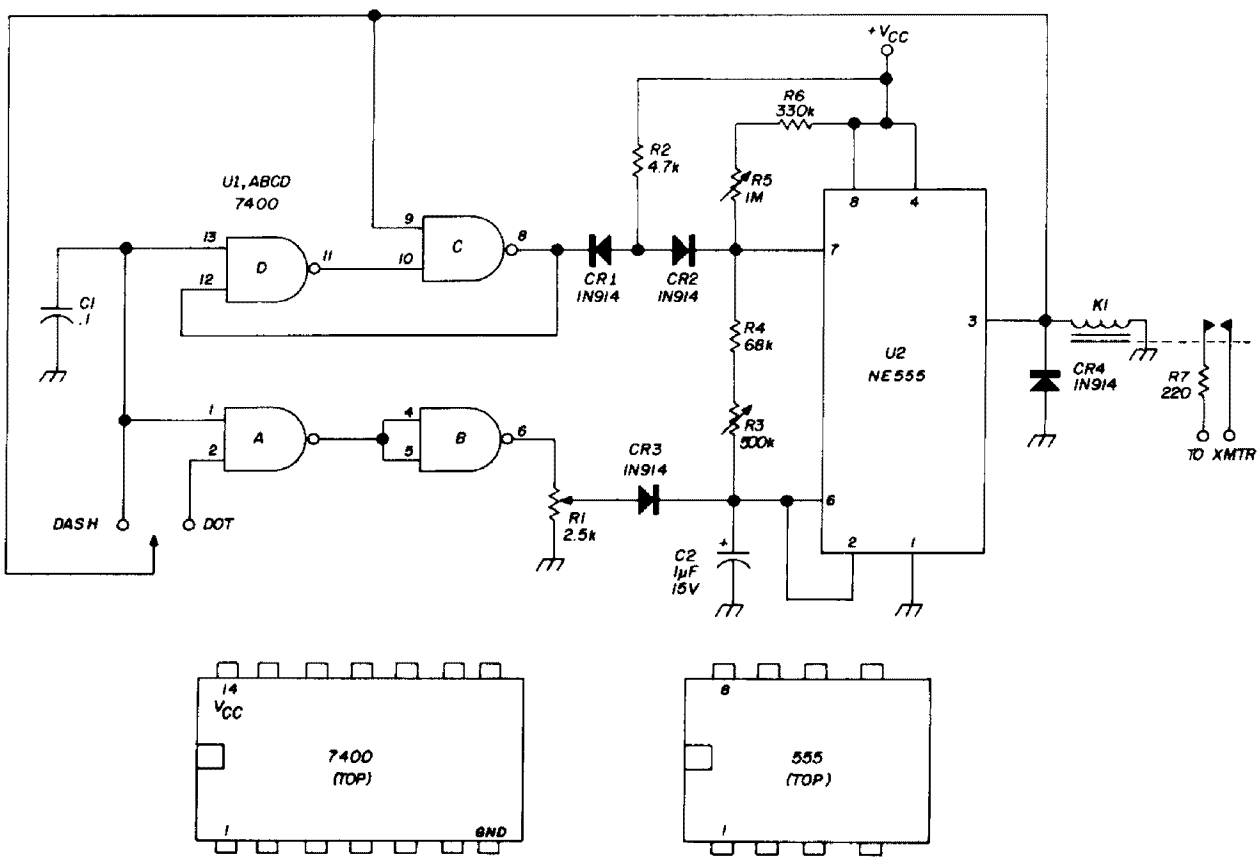


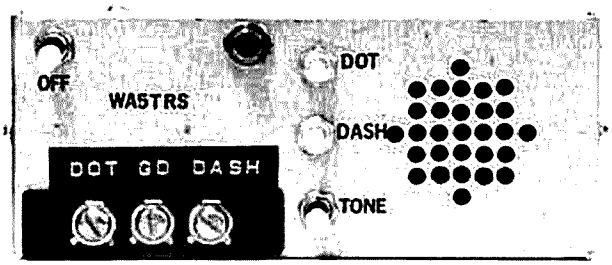
fig. 5. Variable dash length keyer circuit. Relay K1 is a 1A5AH, manufactured by Electronics Application Company, 2213 Edwards Avenue, S. Elmonte, California 91733.

input of gate A, causing its output to go high. Gate B inverts this to a low which allows the bias on pin 2 of U2 to be removed. The charge on C2 is bled through R3 and R4 to pin 7 of U2, which is grounded. When the voltage on pin 2 drops to 1/3 Vcc, the internal flip-flop toggles, opening the ground on pin 7 and presenting a high state to pin 3, the output.

The states of gates A and B reverse but U2 is not affected by the bias change because C2 is charging above the bias

and B would have again changed state at the beginning of the space, removing the bias from CR3 and allowing pin 2 to go to 1/3 Vcc, which would restart the cycle and initiate another dot. If the dot key is opened after the dot is initiated, the bias from R1 will remain on CR3 preventing pin 2 of U2 from dropping to 1/3 Vcc and leaving U2 in an off state.

If the output of U2 is low when the dash key is closed one of the inputs of gate A is forced low, which causes the outputs of gate A to go high and gate B



Variable dot- and dash-length keyer has built-in sidetone monitor. Variable controls in center of front panel control dot and dash length.

to go low. U2 triggers and the output goes high. C1 delays the high signal to the dash paddle for a sufficient length of time to allow gate C to switch low, latching the combination of gates C and D for the duration of the dash. The output of gate C, which is low, causes current flow through R2 and CR1 and a resulting low condition on CR2.

CR2 now charges through R3, R4, R5 and R6 until pin 6 of U2 reaches 2/3 Vcc. At that time the flip-flop toggles, ending the dash and beginning the discharge of C2. C2 discharges through R3, R4 and pin 7 of U2, resulting in the length of the dash being a function of the values of R3, R4, R5 and R6, and the length of the space being a function of the values of R3 and R4.

If the dash key has been kept closed another dash will be initiated at the end of the space when the output goes high. It is impressed on one of the inputs of gate C, latching it for the duration of the second dash. If the dash key is opened after dash initiation, gates C and D will reset when the output of U2 goes low.

adjustment

The standard duration of a dash is three times the duration of a dot. Duration of a space equals that of one dot. Adjustment of the keyer is best accomplished by setting the dot speed first and then the dash length. Close the dot key and count the dots generated in 2½ seconds (or 5 seconds, dividing by 2). This will be close to the words per minute rate. Set R3 to the desired speed. Now, close the dash key and adjust R5 until the number of dashes generated in 5 seconds equals the number of dots generated in 2½ seconds. (One dash plus one space equals 2 dots plus 2 spaces.) The keyer is now set correctly. If, at extremely low or high speeds, it is desired to vary the dot-dash length ratio from the standard to improve readibility, controls R3 and R5 can be adjusted to suit the individual operator's taste.

sidetone generator

On either keyer it is desirable to include a tone generator to allow tune up without keying the transmitter sidetone to allow code practice with the keyer or to simply enable "show and tell" demonstrations. There is nothing exciting about a tone generator, but the one I used in these keyers is about as simple (and cheap) as one can be.

Fig. 6 shows the tone generator circuit using, you guessed it, another NE555 timer. In this circuit, the NE555 is wired as an astable multivibrator (which means oscillator) the same as the dot generator on the keyers. Refer to the operation of U1 on either keyer for description of why it oscillates. The only difference in hookup is pin 4 of U3. This is a reset pin. Grounding of this pin (less than 0.7 volt) forces pins 7 and 3 to logic low, regardless of what else is going on. When the keyer is making a dot or dash, the reset pin is high and the timer is allowed to cycle and generate a tone. Normally, the

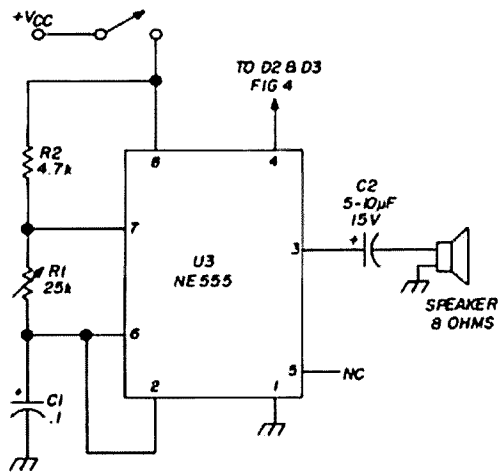
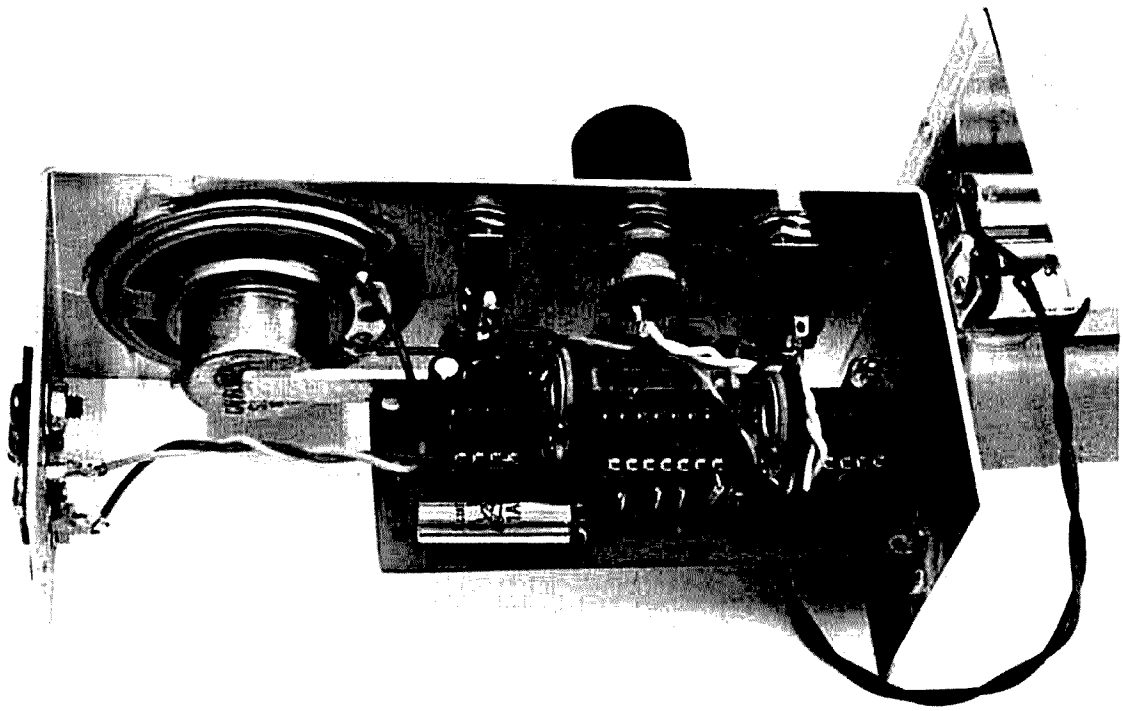


fig. 6. Sidetone generator circuit. Adjust value of C2 for suitable volume.

output of pin 3 would be a square wave but because of reactive loading by the 8-ohm speaker, the output looks like a sloppy sawtooth. However, the resulting tone is satisfactory.

Frequency is adjusted by R1 to suit the user. Since the timing calculations are upset by the speaker load the value of R1 was determined experimentally. R2 is only included to prevent pin 7 from drawing unnecessary current when it is grounded. C2 was included to limit the current which would flow each cycle after the initial voltage rise has thumped the speaker cone and done its job. The value of C2 can be varied to achieve the desired volume level.



Construction of the simple IC electronic keyer. A reed relay is used to key the transmitter.

construction

I chose to use prepunched, unclad circuit board with holes on 0.1-inch centers. Conductor paths and solder connections were made by combining adhesive backed conductors (Circuit-Stik or equivalent) and bus wire insulated with teflon spaghetti. The integrated circuits were plugged into Molex connectors.

When using this method of construction, first decide on your layout, then stick on the conductors for the IC and other components. When this is done install the Molex connectors and solder them in place. Then break off the tie strip between connectors, install the ICs and start wiring.

The ICs are installed before wiring so the Molex connectors don't fall out of the perf board when you're soldering nearby. The tie strip could be left intact until the wiring is finished, but to test for small solder bridges between pins the tie strip must be removed to check for short-circuits with an ohmmeter.

power supply

I prefer a power supply that doesn't have to be plugged into the wall, does not require recharging or battery replacement, and doesn't run down when it's accidently left on — unfortunately, it hasn't been invented yet. Therefore, I ran one keyer with four penlight batteries and the other with a regulated power supply.

The fixed-ratio keyer was battery powered and ran satisfactorily between 4.5 and 7 volts, the range to be expected from batteries. The bias voltage on U1 was adjusted on the threshold of free running with the speed control set to maximum and supply voltage at the minimum which would operate the relay. This should be around 4.5 to 4.7 volts, depending on the relay and ICs obtained.

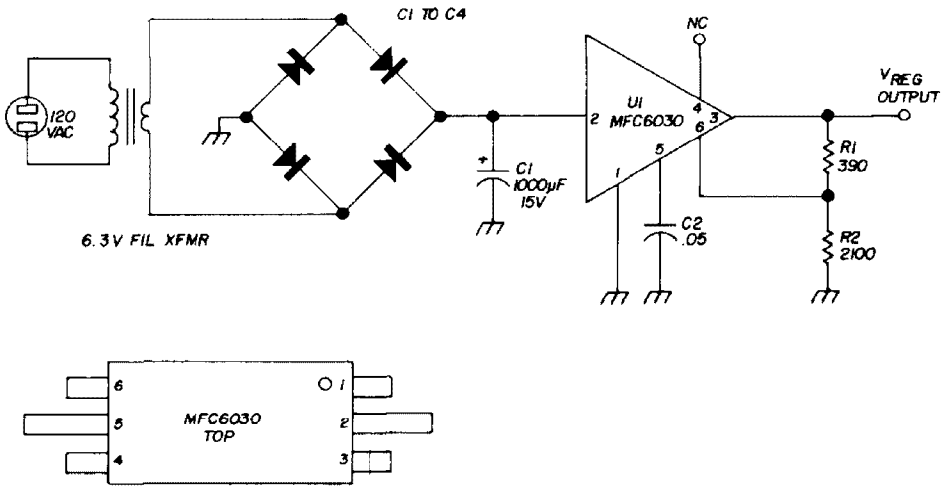


fig. 7. Regulated 5-volt power supply for the NE555 electronic keyers. Diodes CR1-CR4 are 500 mA, 50 PIV. See text for computing values of R1 and R2.

The keyer, thus adjusted, has a built-in warning that the battery is getting low. While making a character, current drain is 46 mA with sidetone on, 35 mA with it off. Quiescent current is 27 mA.

I tried a timer circuit which interrupted the power if the keyer was left idle for ten minutes, but it drew valuable battery power and added unnecessary complexity to the keyer. The regulated power supply in fig. 7 is good for up to 200 mA if the transformer and filter capacitor are stout enough. R1 and R2 establish regulated output voltage and the values may have to be determined experimentally. The reason for experimenting is the manufacturing tolerance of an internal voltage reference in the MFC6030. The published specifications state a range of 3.8 to 4.8 volts. Every one of several I have used measured 4.2 volts.

This can be measured by wiring the rectifier and filter circuits, installing C2, and connecting pins 2 and 5 to V+ and ground, respectively. Temporarily connect pins 6 and 3 together and measure the voltage from this connection to ground. This is the internal reference voltage. The supply voltage must be at least three volts above this for the circuit to function properly. In fact, the input voltage for five volts output should be at least eight volts. The maximum is limited by package dissipation of watt. (Voltage drop X load current). The formulas for computing R1 and R2 are

$$R1 = \frac{R2\,(V_{reg} - V_{ref})}{V_{ref}}$$

$$R2 = \frac{Vref}{I_{ref}}$$

Where V_{reg} is the desired regulated voltage, V_{ref} is the desired reference voltage on pin 6, and I_{ref} is the current flow through R1 and R2 (2 mA minimum).

For example, with a 5-volt regulated output and 4.2-volt reference

$$R2 = \frac{4.2 \text{ volts}}{2 \text{ mA}} = 2100 \text{ ohms}$$

$$R1 = \frac{2100\,(5.0 - 4.2)}{4.2} = 400 \text{ ohms}$$

Since 400 ohms is not a standard value,

*EI Instruments, Inc., 61 First Street, Derby, Connecticut 06418.

choose the next lower value, 390 ohms. Normal component tolerances should cause no problems. However, the output voltage should be checked to make sure it falls between 4.8 and 5.2 volts.

final comments

Within the pages of this magazine are the means to generate the thrill of discovery in anyone with a little curiosity and even less money. The cost of the parts is certainly nominal; the 7400 quad two input NAND gate cost me 26c, the 7470 JK flip-flop, 42c, and the NE555, 98c. At those prices I'm not afraid of blowing something up. The devices must be pretty sturdy because I make a lot of mistakes and haven't burned up an IC yet. I would recommend, for the sake of spontaneity, the use of a breadboard such as the one manufactured by EI Instruments, Inc.* I have wasted a lot of perf board before I got smart. Now I can wire up any of the circuits described here in less than fifteen minutes (including mistake corrections).

I feared that rf would create many problems without shielding and bypassing these circuits. Actually, there was no problem around my equipment with the transmitter on but there was a problem with stray, power-line field-generated voltage, and transients carried by the power line to my home. A large metallic tool held in the hand and touched to the circuitry would trigger the timer. I could not duplicate the condition in the lab where a storage oscilloscope was to be used to track down the method of spurious triggering. The final solution, on my radio bench at least, is to refrain from touching the live circuitry with large metallic tools!

references

1. Ed Noll, W3FQJ, "Circuits and Techniques," *ham radio*, March, April, June and July, 1972.
2. Signetics Data Sheet, Timer NE555, Signetics, 811 East Arques Avenue, Sunnyvale, California 94086.
3. Don Aldridge, WA7RLL, "The Micro-TO Keyer with TTL ICS," technical correspondence, *QST*, September, 1972.

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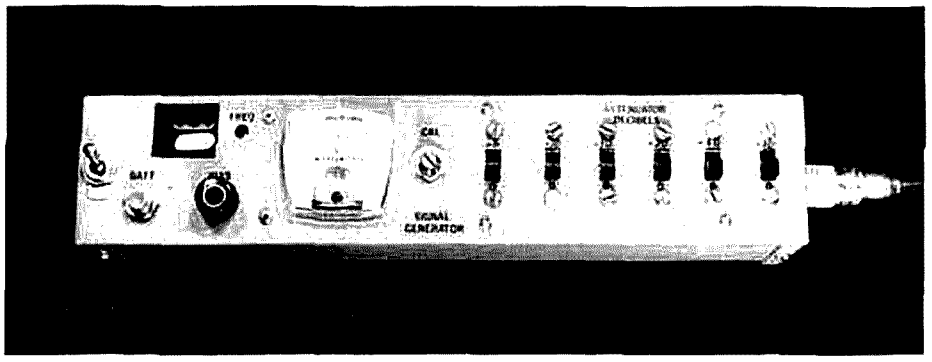
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crystal test oscillator and signal generator

This stable crystal test oscillator uses one low-cost IC to provide a stable rf test signal from 1 to 10 MHz

A.A. Kelley, K4EEU, 2307 South Clark Avenue, Tampa, Florida 33609

This test oscillator was built to fill the need for a stable, resettable test signal source for aligning communications receivers. A handful of crystals gives several accurate marker frequencies with no tuning adjustments. The unit also serves as a handy crystal tester with a built-in meter to indicate relative activity.

The circuit uses a single Motorola MC799P integrated circuit that costs about $1.00, but will oscillate with any crystal from approximately 1 MHz to 10 MHz. The basic circuit is stable and has been used with a frequency counter,[1] but slightly smaller capacitor values have been used here to insure that less active crystals will oscillate. The circuit provides 32 pF crystal loading; a small E.F. Johnson variable trimmer capacitor may be used to adjust the crystal to the exact frequency. This capacitor may be omitted if you don't need this feature.

The circuit is not critical so use the parts you have on hand. It will even oscillate with 450-kHz crystals *if* the 22-pF series capacitor is increased in value. The variable bias pot is broadly adjustable to compensate for battery voltage.

construction

The oscillator is built inside a 2½ x 2¼ x 12-inch Minibox *with* crystal sockets submounted so that some shielding is provided by the surrounding box. If the generator were to be primarily used at low output levels, it might be advisable to provide a cover plate over the crystal socket area. Two sockets for HC6U and FT-243 will provide for most types in use today.

The six attenuator pads are constructed with standard resistor values and mounted on dpdt slide switches obtained from Weinschenker.* Shields fabricated from scrap aluminum are closely fitted between switch sections to control leakage across the pads, and the interior coax fitting is shielded to prevent leakage around the assembly.

The inexpensive Japanese 50-microampere meter serves as a battery tester and gives instant indication of crystal activity. The output level may be set for half-scale on the meter to monitor the input to the attenuators.

After construction, my supply of surplus FT-243 crystals was checked and several were rejected. It was found that crystals defective in this checker would not oscillate in tube oscillators either. A few could be salvaged by careful cleaning.

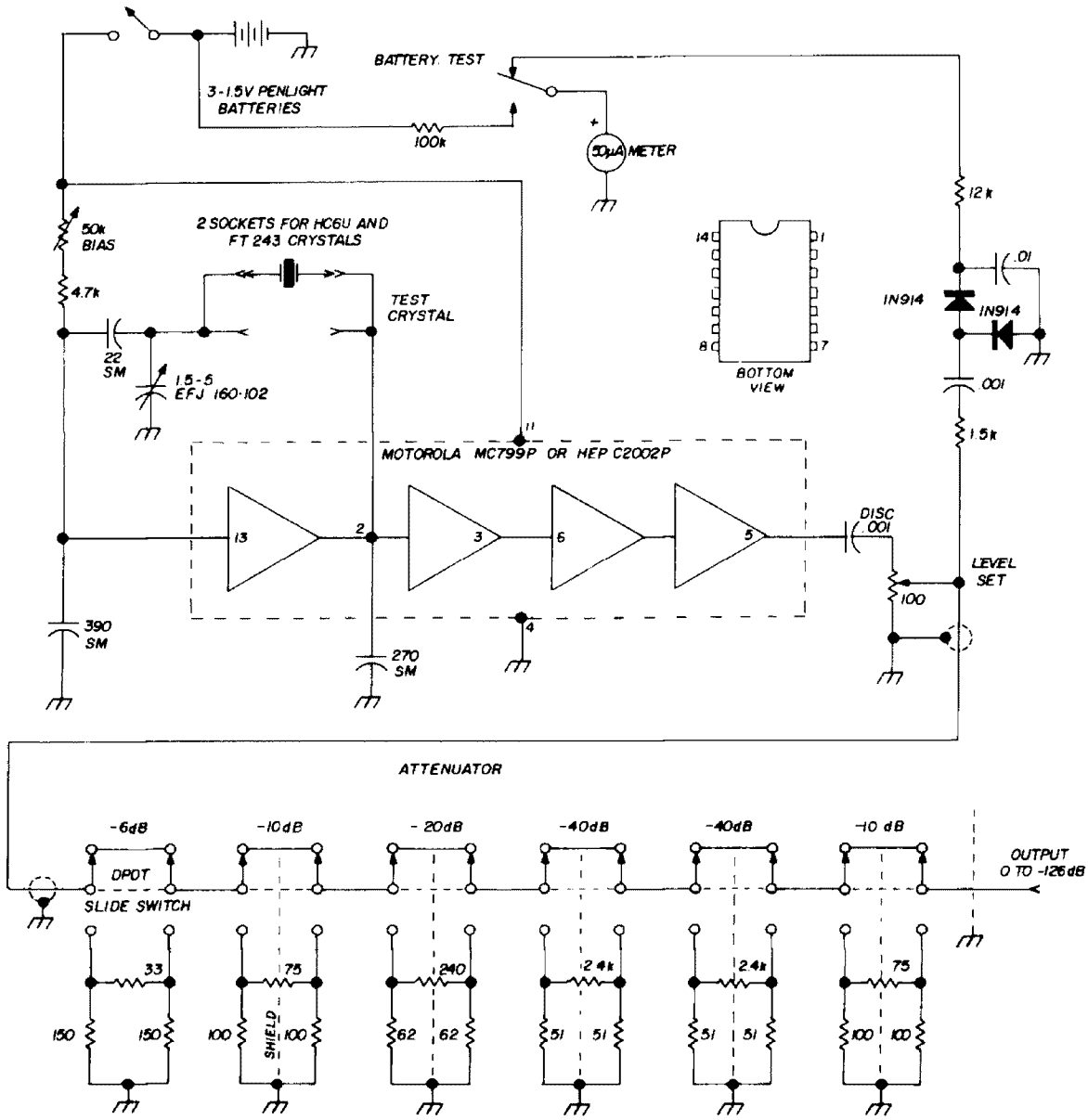


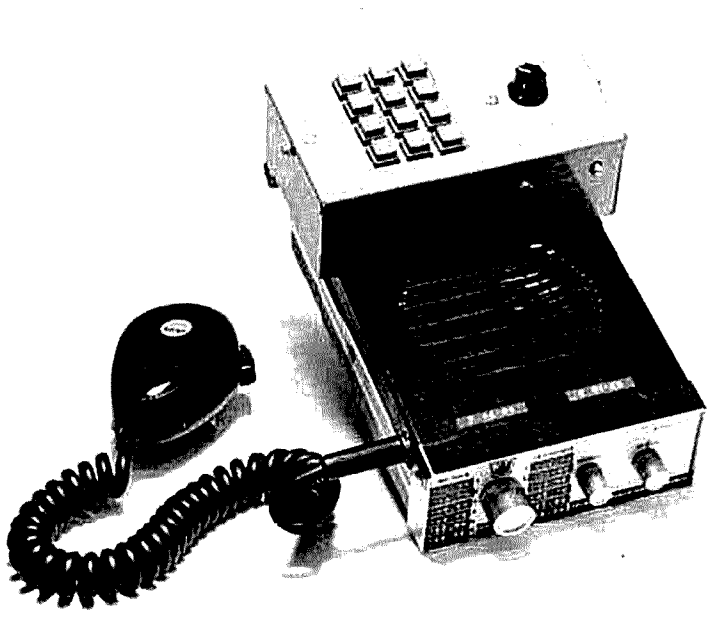
fig. 1. Circuit diagram for the crystal test oscillator and signal generator. MC799P IC will oscillate with any crystal from 1 to 10 MHz. Output attenuator uses standard resistor values to provide up to 126 dB of output signal control.

*M. Weinschenker, Box 353, Irwin, Pennsylvania 15642. Slide switches are priced at six for $1.00, dpdt, red or black.

reference

1. A.A. Kelley, K4EEU, "Compact Frequency Counter," *ham radio*, July, 1970, page 16.

ham radio



solid-state mobile touch-tone circuit

This all solid-state Touch-Tone circuit provides automatic, mobile fm operation

Roy C. Hejhall, K7QWR, Motorola Semiconductor Products, Inc.

Since joining the growing crowd of two-meter fm repeater enthusiasts over a year ago, I have observed a variety of telephone Touch-Tone pad interface circuits for connecting the pad to the transmitter. Most have had various features which I felt could be improved upon by an all solid-state version. Thus, I launched the design effort which resulted in the circuit described in this article.

background

One of the better circuits to come to my attention is the one described in an excellent article by WØLPQ.[1] This circuit has several advantages over other Touch-Tone interface circuits commonly in use: automatic keying of transmitter with delayed drop-out, automatic connect-disconnect of Touch-Tone audio output to transmitter audio system, and no transformer or battery required.

It was decided that the new design would incorporate the above features while replacing the relay with an all solid-state circuit. This approach has the advantage of all electronic circuitry (no

moving parts). Total cost of all four transistors is only $1.56.

I also decided to inject the audio signal someplace downstream in the transmitter speech amplifier, instead of at the super-sensitive, high-impedance microphone input. There is adequate signal level available to do this. Making connection to a higher signal level, lower impedance point in the speech amplifier minimizes hum worries and eliminates the need for shielded wire to carry the audio signal to the transmitter.

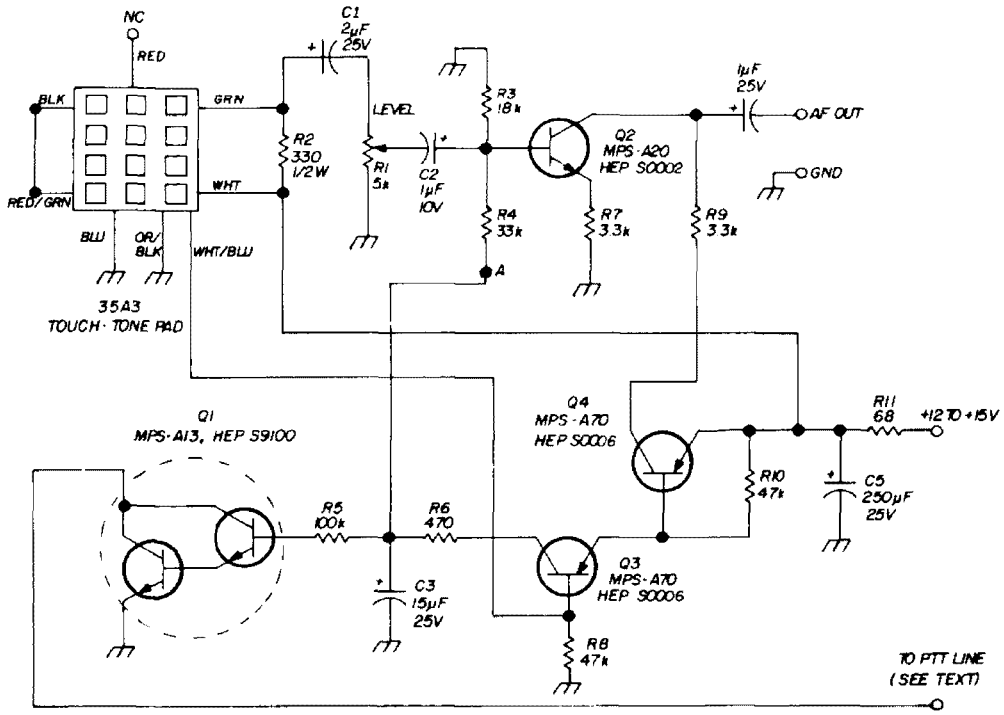


fig. 1. Schematic diagram of the mobile Touch-Tone circuit. The 5.1k resistor connected between the blue and white/blue leads inside the 35A3 must be removed.

the circuit

Fig. 1 shows the schematic of the Touch-Tone interface. The Touch-Tone pad I used is a Western Electric Model 35A3. All wires on the 35A3 are color coded and the schematic indicates the connections by wire color.

One modification of the 35A3 is required. Remove the 5.1k resistor which is connected to the white and white-blue leads. This is easily located and snipped out.

The green wire is the audio output lead of the 35A3. The signal goes through level control R1, unity gain amplifier Q2, and on to the transmitter.

Transistor Q3 is a dc switch and Q4 is both a dc and signal path switch. Q1, a Darlington-connected transistor pair in a single package, performs the function of the transmitter push-to-talk switch.

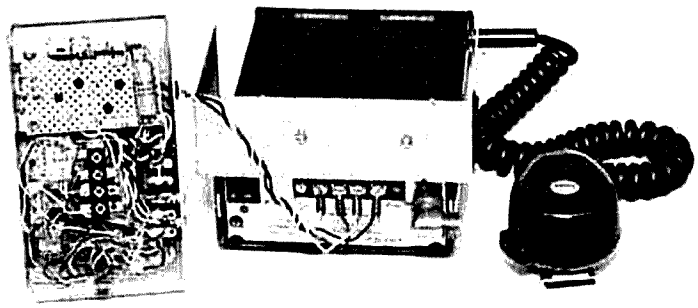
All four transistors are normally *off*. When any button on the 35A3 Touch-Tone pad is depressed, Q1, Q3 and Q4 become essentially short circuits while Q2 is biased *on* to perform its amplifier function.

The automatic audio disconnect is accomplished by turning transistors Q2 and Q4 off. Under these conditions, essentially an open circuit is presented to the point of signal connection inside the transmitter. Transistor Q3 performs the function of turning Q1 and Q2 on and off. The dc input filter R11-C5 was added to reduce spikes (or ac ripple when operating from an ac supply).

I liked WØLPQ's automatic keying and delayed drop-out features, and incorporated similar functions in this circuit. This results in the transmitter being keyed automatically when any 35A3 button is

depressed. It remains keyed until after the last digit of the phone number is dialed, instead of switching back to receive between each digit. This delay is accomplished by the gradual discharge of C3 through R3, R4, Q2, R5 and Q1 after Q3 is switched off. Resistor R6 limits the turn-on charging current through Q3 to a safe value.

There is nothing critical at all about the circuit, and layout is left to the constructor. In fact, as one who works with rf most of the time, building a noncritical audio circuit is a refreshing change of pace. A printed-circuit is available to those who are interested.* Level control R1 should be a "set it and forget it" control, so it may be an internal screwdriver adjustment.



Bottom view of the Touch-Tone circuit. At top is the circuit, mounted on a 2 x 3" piece of Vector perf-board. Touch-Tone pad is below. The four terminals on the back of the HR-2A were modified to permit connection to the solid-state Touch-Tone circuit.

connection and checkout

The dc feedline is connected directly to the automobile electrical system with no voltage regulation or additional filtering beyond that shown in fig. 1. The circuit also functions very well when connected to an ac supply with 1.2 volts peak-to-peak ripple when the transmitter is keyed. If your car is noisy, the value of capacitor C5 may have to be increased.

The lion's share of total current drain is the 16 mA or so drawn by the 35A3 Touch-Tone pad. The remainder of the circuit draws less than 2 mA when on and less than 0.5 mA when off.

The automatic transmitter keying circuit is designed to be connected in parallel with the PTT mike switch in the Regency HR-2A. Before making connection to the PTT line of other rigs, the following must be determined:

1. The PTT circuit must be similar to that shown in fig. 2.
2. The PTT line current must be less than 300 mA with up to 16 Vdc input to the rig.

If your rig does not meet these requirements, the automatic keying portion of the interface circuit must be redesigned, not connected, or eliminated.

Not connecting this function simply means leaving the collector of Q1 open. To eliminate this portion of the circuit, delete Q1, R5, C3 and R6; connect the collector of Q3 to the lower end of R4 (point A). Neither of these options will have any effect on the operation of the remainder of the interface circuit.

Advantages of connecting the audio output someplace downstream in the transmitter speech amplifier were discussed earlier. However, for the builder who prefers making the connection to the microphone input, the circuit should function just as well that way with no modifications. In the HR-2A I connected the audio output of the circuit to the junction of C226 and R223 between the second and third stages of the speech amplifier.

Excessive rf from the transmitter can

*Printed-circuit boards are now available from Contact, Inc., 35 West Fairmont, Tempe, Arizona 85281, for $3.00 for the board only, or $11.50 fully wired and tested, plus 25 cents postage and handling. Please direct all correspondence regarding the board to this address and not to the author.

cause the interface circuit to malfunction. To prevent this I bypassed the PTT and dc supply lines inside the HR-2A. This was accomplished by connecting 0.001-μF disc ceramic capacitors between the lines and ground at the point where the lines leave the HR-2A cabinet to go to the Touch-Tone circuit.

The only adjustment is R1. If deviation measuring equipment is available, set it for the deviation specified for your local auto-patch system. Otherwise, set R1 for reliable operation of the patch.

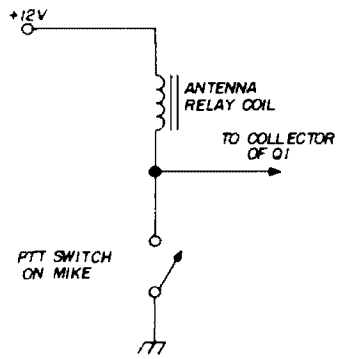


fig. 2. Push-to-talk circuit for the Regency HR-2A vhf fm transceiver.

circuit variations

The time that the transmitter remains keyed after release of the last button is determined by the capacitance value of C3. To shorten the time, decrease the capacitance; to lengthen it, increase the capacitance.

A temporary disabling mode for the automatic keying feature may be included by adding a switch from the base of Q1 to ground. Closing this switch will prevent automatic keying of the transmitter without affecting the operation of the remainder of the circuit.

acknowledgment

Special thanks are due to Dick Evans, W7BBW, who constructed and field tested the prototype unit.

reference

1. William P. Lambing, WØLPQ, "Mobile Operation with the Touch-Tone Pad," *ham radio*, August, 1972, page 58.

ham radio

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100 Mfd	@	15 Volts	5 for	$1.00
250 Mfd	@	30 Volts	3 for	$1.00
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	1N270	DIODES	20¢ ea. ppd.

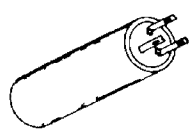
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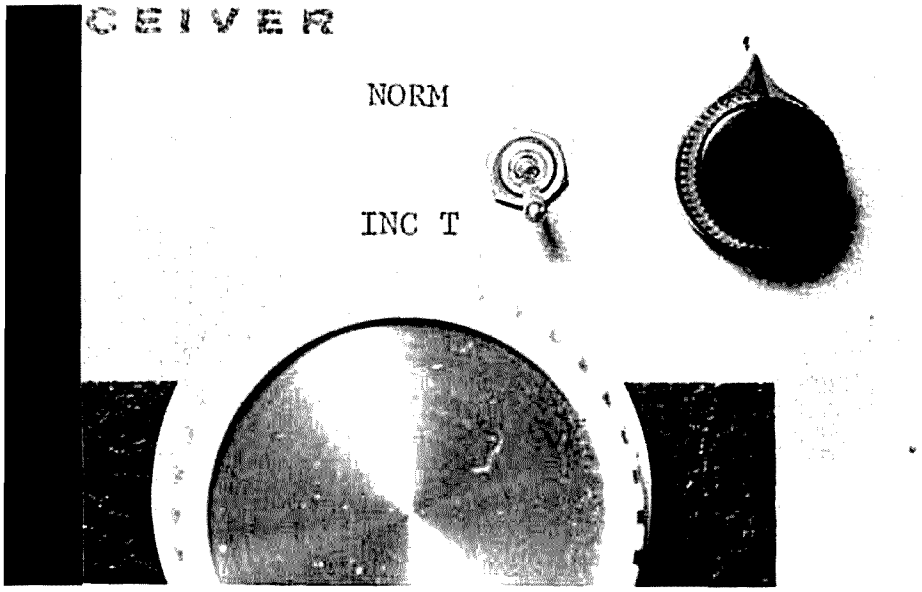
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HW16 modifications

for vfo operation

Easily made mods to enhance the versatility of this popular rig

James T. Takashima, WB6MZN, 19104 Wadley Avenue, Carson, California

The Heath HW16 transceiver is very popular with many Novice-class hams. However, the HW16 transmitting frequency is crystal controlled, so the owner of this equipment will probably be buying a new rig or a vfo now that the FCC has lifted the crystal-control restriction for the Novice-class licensee.

The HW16 can be easily modified to incorporate vfo capability. This and other desirable features are described, which will enhance the versatility of the HW16. The vfo mod alone will cost about $5, and the options will cost another $9 or so, even if all parts must be bought new.

modifications

The modifications are indicated with heavy lines on the block diagram, fig. 1. I believe this is a simple and logical way to receive and transmit on frequencies controlled by the vfo built inside the rig.

The HW16 heterodynes the incoming signal with three oscillator frequencies to produce the audio output when receiving. So let's up-convert these oscillator frequencies when transmitting and answer that CQ right on frequency! This "reversed double conversion" is accomplish-

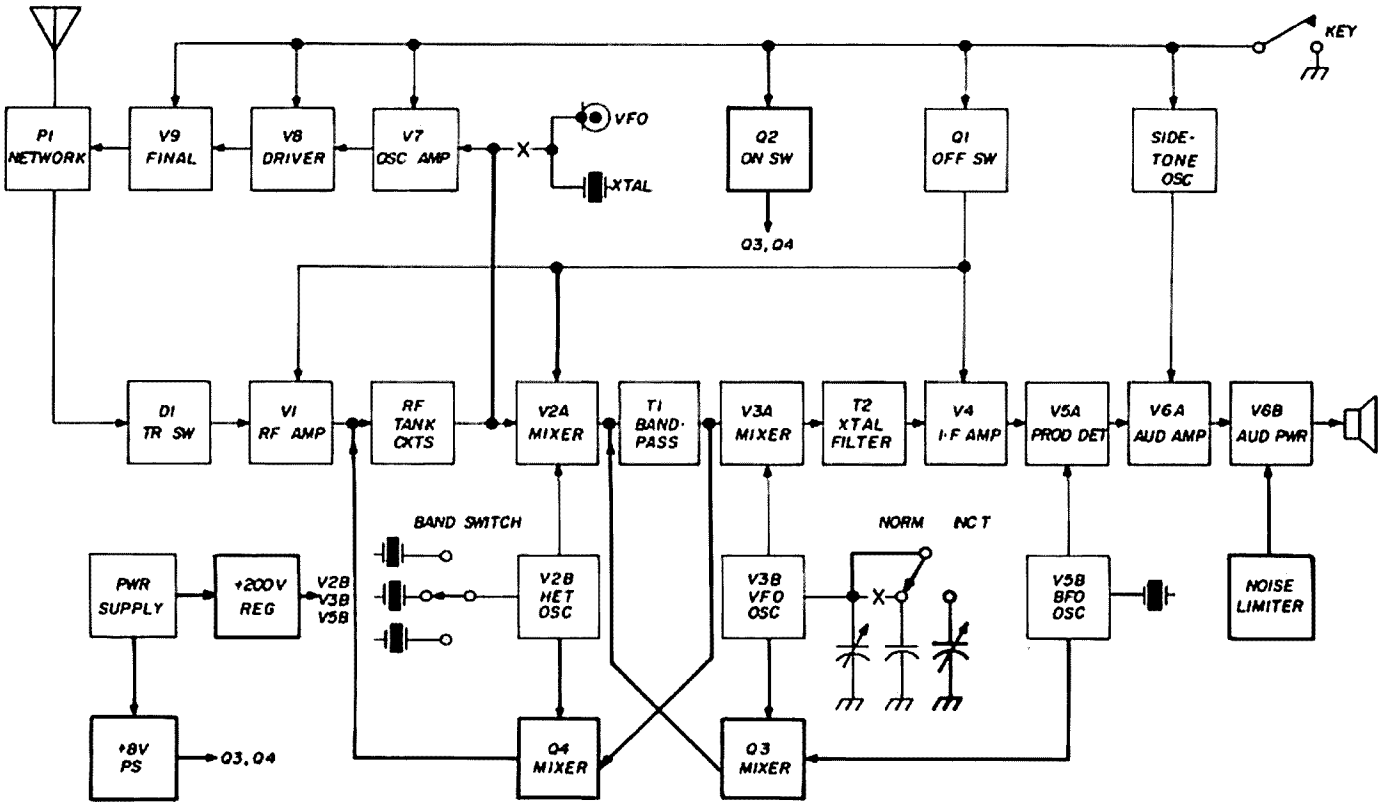
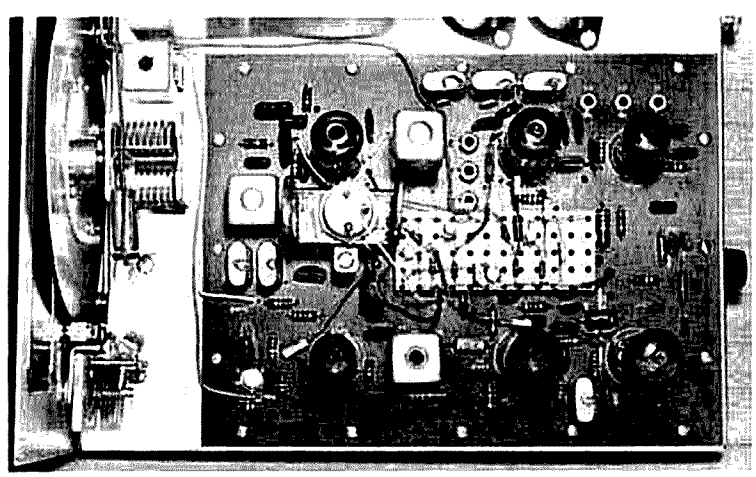


fig. 1. HW16 modifications for improved versatility. Modifications are shown in heavy lines in this diagram and in the schematic, fig. 2.

ed using the oscillators and tuned circuits already in the HW16 for all bands. The rf signal thus generated is at the frequency to which the receiver is tuned (minus the audio frequency, to be exact). It is then amplified through the three stages of the transmitter.

construction

Heavy lines and symbols in the schematic, fig. 2, show all the additions and changes needed for the modification. The transistor mixers work well when their base-current limiting resistors are chosen for the collectors to operate at about +4 Vdc. Other type transistors with equivalent current gain and cutoff frequency should also make good mixers. Power and voltage requirements are extremely low. The mixers and ON switch are on a 1 x



New components are mounted on a Vector board located between V2 and V5. Existing Heath components have been mounted on the other side of the existing circuit board to provide clearance for modifications.

3-inch Vector board mounted between V2 and V5, where existing components have been mounted upside down on the other side of the printed circuit board to clear the area. The change of R24 resistance value is to increase the receiver sensitivity, and the R28 change is to improve oscillation stability of V2B. A quency change is about ±10 kHz per ±2-pF change of the 6-pF variable capacitor. The switch is placed in NORM position when this mode is not used.

Other optional features are supply voltage regulation for the oscillator tubes to prevent possible chirping and the noise limiter to eliminate key clicks in the side

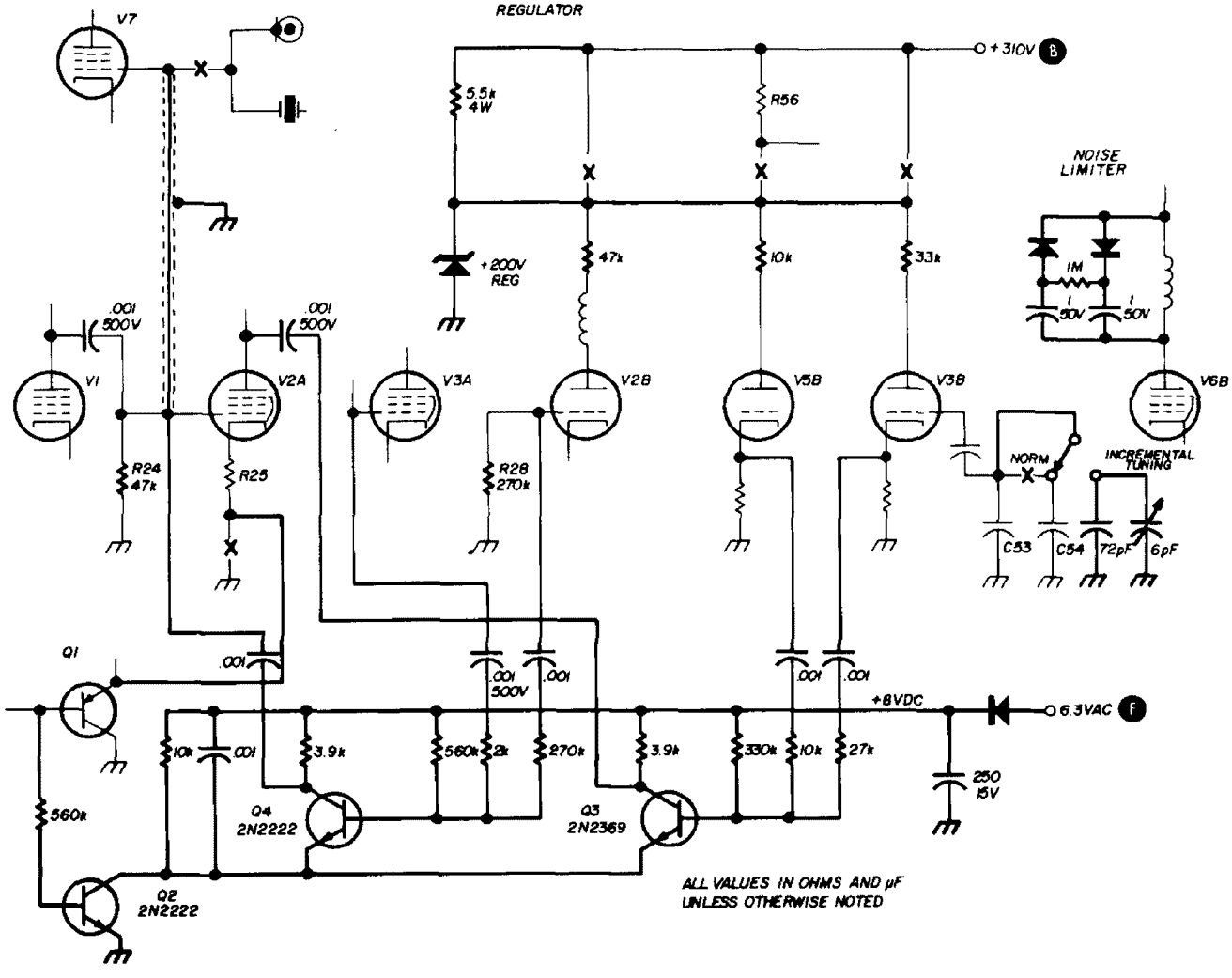


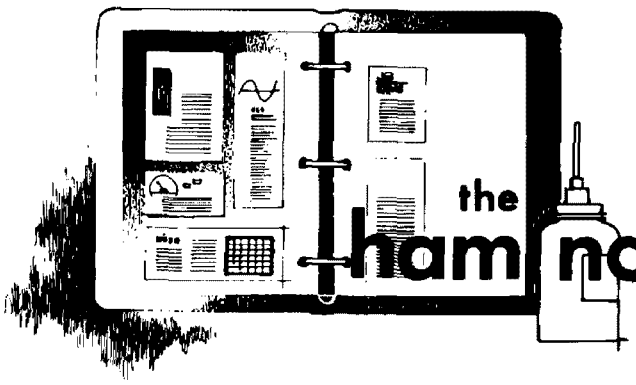
fig. 2. HW16 schematic showing modifications. Features include transmit vfo, incremental tuning, power-supply regulation for the oscillator tubes, and side-tone key click filter.

small amount of retuning of tank circuits may be necessary.

other improvements

Incremental tuning is an optional feature to provide an extra tunable receiving capability around a transmitting frequency, or vice versa. The switch and variable capacitor for incremental tuning are mounted on the front panel — the only externally visible modification. Fretone. The zener diode shown in the schematic is actually the collector-base portion of a silicon power transistor on a good heat sink. Tube-type voltage regulators and different transistors can be used, but all the parts for the modification are from junk boxes in my case. Tubes (6CB6 for Q3 and 6U8 for Q4) were used initially and successfully except for much greater heat dissipation.

ham radio



the ham notebook

tone-burst generator

The tone-burst generator circuit shown in **fig. 1** uses a new, low-cost Signetics IC, the NE555. Although this circuit was designed specifically for the HR2 vhf fm transmitter, the circuit could be easily adapted to other vhf fm equipment. The 5-megohm audio pot in the U1 circuit provides a nice scale expansion of burst length in the 0.3- to 1-second range. Tone frequency is controlled by the 25k linear pot in the circuit of U2; output frequency is from approximately 1300 Hz to 2500 Hz.

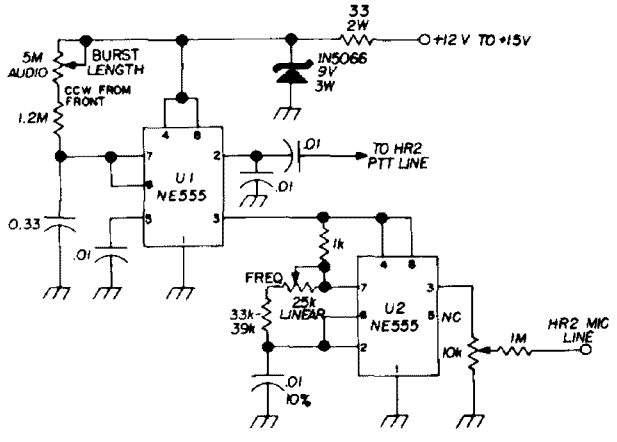


fig. 1. Circuit for the IC tone-burst generator.

There are many other circuit possibilities for the new Signetic NE555 IC, including tone generators, timers, repeater control, AFSK modulators, code oscillators, pulse detectors, clock generators and electronic keyers.

Phil Elrod, K4COF

zulu time

Frequently, someone makes a remark over the air when using *Z* after Greenwich Mean Time, wondering why that letter is used. For many years, the Military services have used letters, except for the letter *J* (nobody seemed to want Jig Time), to designate time zones. Eastward from Greenwich, the zones are lettered A through M (except J). Westward, the letters N through Y apply.

Thus, within the United States, the following apply for 0000Z (GMT):

EDST	2000Q
EST and CDST	1900R
CST and MDST	1800S
MST and PDST	1700T
PST	1600U

Note that M and Y both apply to the zone across the International Date Line; the time is the same, but the date is different. M applies to the west side of the date line whereas Y applies to the east side.

Bill Conklin, K6KA

simple timer

The article in the September, 1972, issue of *ham radio,* on simple repeater-control timers prompts me to point out a useful integrated circuit which I have used in several construction projects — the Signetics NE555V timer module.

The NE555V is an 8-pin, half DIP package selling for about $1.00 that can be used as a monostable or a stable multivibrator with a timing period from 2 microseconds to an hour. Further applications include missing pulse detection, pulse-width modulation, pulse-position modulation or voltage-controlled multivibrator.

A chief advantage of the NE555V over other IC one-shots is the ease in obtaining long time delays. The 74 series IC one-

shots provide time delays given by T = 0.32 RD, where R is restricted to 30 to 50 Kilohms. The NE555V provides time delays given by T = 1.1 RC, where R can be as high as 10 megohms. Thus, with the same RC values, the NE555V gives a supply approximately 9 volts dc. This is connected to an NPN emitter-follower voltage-regulator circuit. A 4-volt zener in the base of this transistor provides a regulated 3.3 volts at the emitter.

The 120-volt loop supply is added by

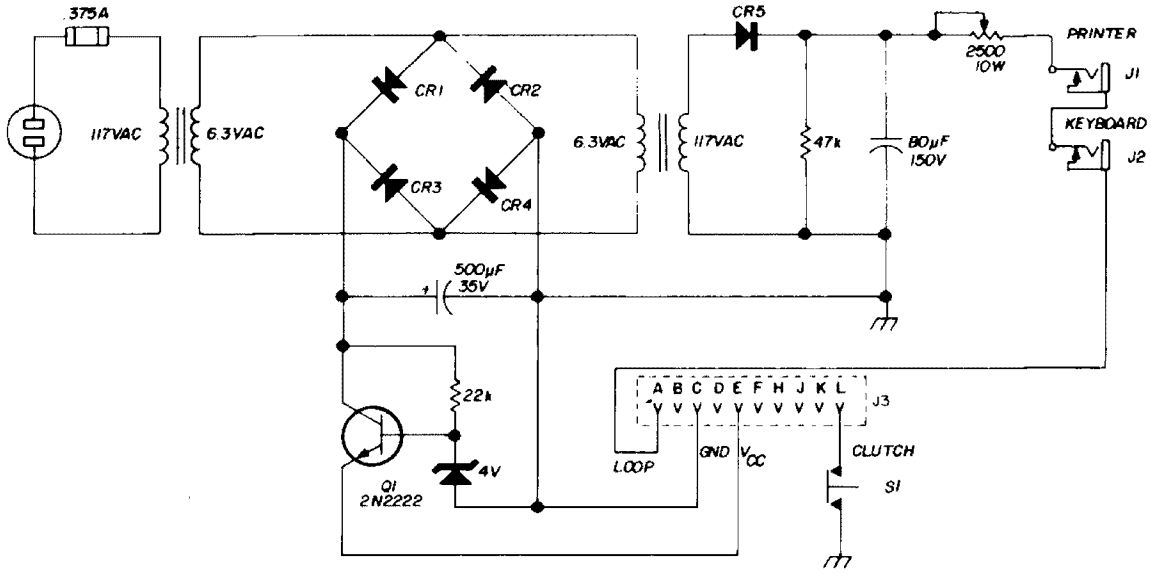


fig. 2. Circuit for the RTTY test generator. Diodes CR1-CR5 are 50 PIV, 100-mA diodes. S1 is a normally closed, momentary contact switch. Phone jacks J1 and J2 must be insulated from the chassis.

factor of three advantage in delay time, while the large allowable resistance provides delays up to an hour.

A word of caution is in order when using large capacitor values. If poor quality capacitors are used, the leakage current should be considered in computing the time delay since it acts as a parallel resistance. I have used 1-meg resistors and 150-microfarad tantalums for 3-minute delays with no problems.

Elmer Mooring, W3CIX

RTTY test generator

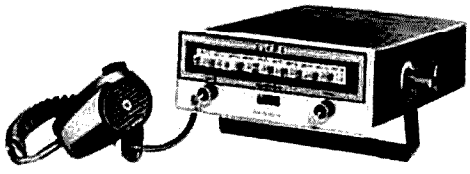
The RTTY RY generator described in the March, 1971, issue of *ham radio* can be made into a compact test unit, including a 120-volt loop supply and 3.3 volt Vcc for the generator board, and housed in a 2-1/8 x 3 x 5¼-inch Minibox. The supply for the generator board is obtained by using a 6.3-volt transformer (Radio Shack 273-050) and bridge to connecting another 6.3-volt transformer back-to-back with the one used in the 3.3-Vdc supply. This provides 110 Vac, isolated from the power line, which is rectified to provide the 120-Vdc loop voltage. A 2500-ohm adjustable resistor permits setting loop current to 60 mA. Two closed-circuit phone jacks are connected in series with the loop supply. With the printer plug in one jack, loop current can be monitored with a milliammeter plugged into the other jack. Local copy can be generated by plugging the keyboard into the second jack and, of course, when the RY generator is active the loop is keyed to give local RY copy.

The clutch circuit is wired in this unit to a normally-closed momentary-contact switch. This provides a steady 60 mA of magnet current which may be keyed by the keyboard for local copy. Depressing the momentary contact switch will cause the loop to be keyed with a stream of RYs until the switch is released.

Tom Gibson, W3EAG

new products

vhf fm transceiver



The introduction of a new 10-channel 2-meter fm transceiver has been announced by General Aviation Electronics (Genave). The advanced GTX-2 is a lightweight, all-solid-state unit manufactured by a leader in the field of navigation and communications electronics equipment for the aircraft and marine industries. It comes complete with quick disconnect power cable, plug-in microphone, SO-239 antenna connector and mobile mounting bracket, and retails for $249.95.

Push-button frequency selection is a major feature of the innovative GTX-2. It comes equipped with a 146.94 MHz communications channel. The remaining nine channels are available for installation at the factory or by the owner. Crystals are available for $6.50 each.

The high performance capability of the compact and lightweight instrument (5 lbs) is made possible through fully transistorized and integrated circuitry. The extensive use of integrated circuits results in decreased size and weight, and significantly reduced power consumption, while providing maximal power output and reliability. The use of ICs also results in more economical manufacturing processes, which contribute to the extraordinarily low cost of the unit.

The new unit features characteristics usually found only in more expensive radios, including a surprising 30 watts of output power. It is readily adaptable for fixed or mobile operation. A multiposition switch allows setting for long-time low power drain operation. Thus, the radio can be operated for extended periods of time with minimal current usage.

In the receive mode sensitivity is less than 0.5 microvolts for 12 dB SINAD. Images are suppressed more than 45 dB and spurious responses are down more than 50 dB. Selectivity is ±8 kHz. Squelch threshold is 0.5 μV maximum.

The transmitter covers the frequency range from 144 to 148 MHz, and features nominal output power of 30 watts. The output matches standard 50-ohm amateur antennas. Frequency deviation is adjustable to 10 kHz maximum. For more information, write to General Aviation Electronics, Inc., 4141 Kingman Drive, Indianapolis, Indiana 46226, or use *check-off* on page 94.

microtransmitter

Lithic Systems has announced a milestone in monolithic integrated circuits — the world's first radio transmitter on a chip.

Designated the LP2000 *Microtransmitter,* the device produces 100 mW pulse modulated, or 50 mW amplitude modulated at 27 MHz from a high stability, regulated monolithic oscillator using external crystal control. A unique transformerless modulating circuit has been created, with buffering between oscillator and modulated output stages. Rf output power and power drain are externally controlled. The IC also includes a low-level modulation preamplifier/tone-coding generator, internal power supply regulation, and a latching power supply switch which draws zero power from batteries in the "off" condition.

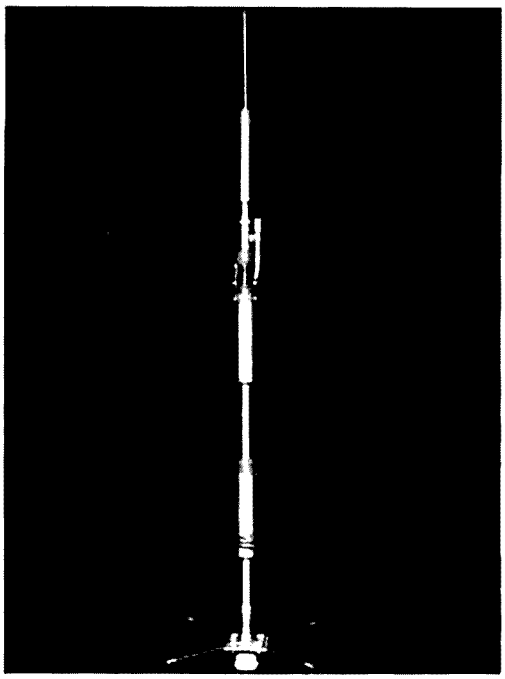
More Details? CHECK–OFF Page 94

The circuit operates from +15- down to +3-volt supplies.

Intended applications include hand-held, mobile, airborne and marine two-way radio; remote controls, and short-range telemetry. Small size, low weight and high reliability make it attractive for biomedical monitoring and security alarms.

The LP2000 is available in a 10-pin hermetic TO-100 package. Small quantity distribution to amateurs and experimenters is through Circuit Specialists, P.O. Box 3047, Scottsdale, Arizona 85257. For more information use *check-off* on page 94.

base-station colinear antenna



The Antenna Engineering B-series base-station colinear antenna is a triple-skirted design operating with a decoupling ground plane. The antenna consists of seven quarter-wavelengths, and is available for all amateur and commercial frequencies in the 140-175 MHz, 220-225 MHz and 420-470 MHz bands. Unlike many antennas of this type, the B-series antenna is at dc ground for positive lightning protection, and the gamma-type feed is located on the radiating structure for symmetrical current distribution. The feed system will match 25 to 100 ohms for use with various transmission lines

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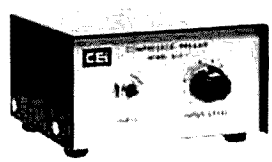
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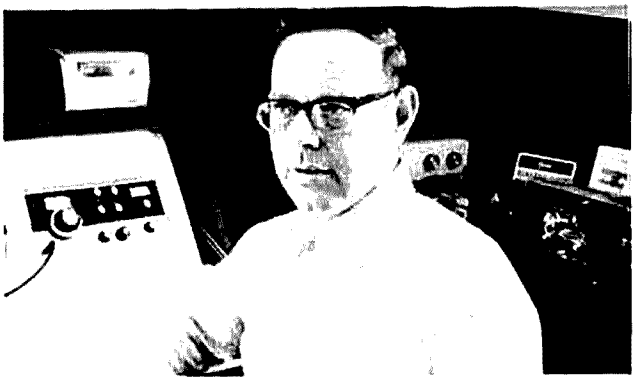
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and in phased-arrays.

The supporting mast is heavy-wall aluminum alloy, and the radials are spring-tempered stainless steel. A mounting receptable is provided for 1-inch NPT pipe. The unit is quite rugged for its light weight. Prices range from $39.95 for the 140-175 MHz version to $29.95 for the 420-470 MHz version.

For more information on the B-series base-station colinear, write to Antenna Engineering Company, Inc., Box 19449, Indianapolis, Indiana 46219, or use *check-off* on page 94.

high-power balun



The new Ultra-Bal 2000 balun from K.E. Electronics is available in either 1:1 or 1:4 ratio models, covers the frequency range from 3 to 30 MHz and features a 2000-watt average power rating. The Ultra-Bal 2000 is weatherproofed by encapsulation in low-loss resin. Unlike other baluns the Ultra-Bal 2000 is totally sealed; there are no drain holes to let moisture enter and cause problems with rf arc-over. The Ultra-Bal 2000 is wound with heavy silver-plated wire for low insertion loss and Teflon-insulated wire provides superior resistance to voltage damage under high swr conditions. The use of Delrin plastic for the center-of-dipole insulator eliminates hardware corrosion, and solid brass output terminals remove dissimilar metal contacts in the amateur antenna system.

The Ultra-Bal 2000 is priced at $8.95 postpaid in the U.S.A., and is supplied complete with instructions, including detailed examples of the various impedance-matching techniques which may be used with the unit. Specify 1:1 or 1:4 impedance ratio. For more information, write to K.E. Electronics, Box 1279, Tustin, California 92680, or use *check-off* on page 94.

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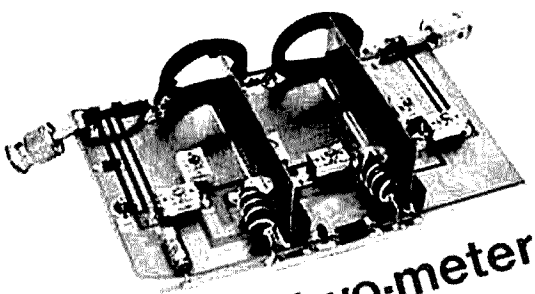
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APRIL 1973



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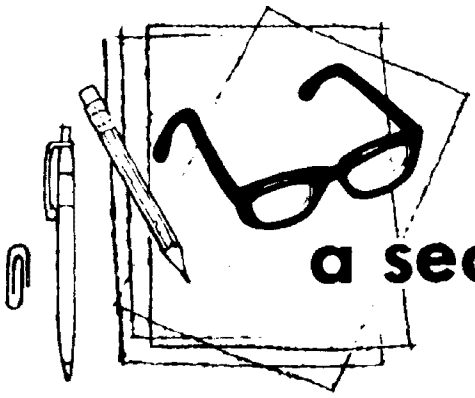
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a second look

by jim fisk

Although any amateur who has worked on the high-frequency or vhf bands knows that geomagnetic storms can raise havoc with radio propagation, new evidence indicates that these same geomagnetic storms are linked to changes in our weather.

Upper air circulations over the Gulf of Alaska, which subsequently affect the weather over the North American continent, tend to become more intense after a period of severe geomagnetic disturbance. Although evidence of this phenomenon was reported over ten years ago, it's only been within the past couple of years that the subject has received serious study by scientists from the National Oceanic and Atmospheric Administration (NOAA) and the University Corporation for Atmospheric Research, both in Boulder, Colorado.

The studies to date have been based primarily on statistical studies — the number of high-altitude storms that follow a period of geomagnetic disturbance as opposed to the number of high-altitude storms following non-geomagnetic storm periods. Although the studies are far from complete, all evidence so far points to increased high-altitude storm activity 5 days to a week after a geomagnetic disturbance. If more can be learned about this storm intensification, it may provide a valuable new tool for increasing the accuracy of the 3- to 5-day weather predictions in the United States and Canada.

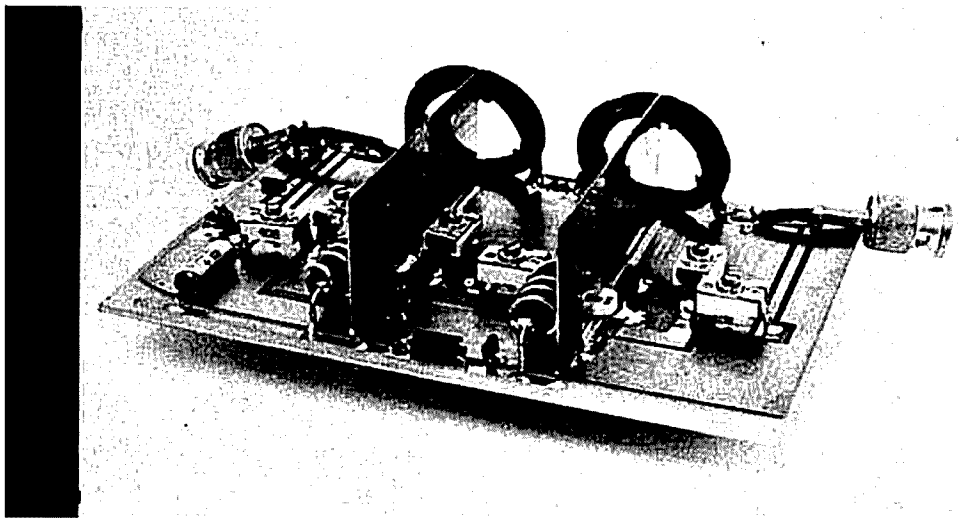
Although the cause and effect of geomagnetic storm intensification is not yet clearly understood, scientists speculate that it may result from geomagnetic modulation of the heat radiated to space from the relatively warm Alaskan Gulf during the winter. This may trigger the formation of cirrus clouds, change the ozone content, and result, possibly, in increased thunderstorm activity.

Satellite infrared data which indicates such changes during the winter of 1971/1972 are now being studied to test the idea of geomagnetic control of cirrus clouds, but further work is necessary. However, the study may provide important clues to the physical process by which the miniscule direct energy inputs from geomagnetic activity in the lower stratosphere can cause such large weather patterns that affect millions of people on the North American continent.

Once the mechanism of this phenomenon has been clearly understood, it suggests a possible way in which man might deliberately (or inadvertently) modify large-scale high-altitude air circulation, and hence, the weather. Controlling the weather, like flying and space exploration, has long been among man's fondest dreams. All it may take is a small amount of man-made magnetic energy, in the right amount and in the right place.

If all this comes to pass I can see it now — the skiiers will be lobbying for snow, the golfers will be lobbying for sunshine, and ecologists will be asking for moderation. More important to amateurs, perhaps, is that if geomagnetics can be used to control the weather, perhaps the same sort of technique can be used to improve long-distance radio propagation during periods of sunspot minimums.

Jim Fisk, W1DTY
editor



solid-state two-meter fm rf power amplifiers

Complete design and construction details for solid-state power amplifiers for two-meter fm

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This article describes three solid-state vhf power amplifiers designed for use on two-meter fm. Since many of the fm transceivers now in use are in the one-watt class, such as the Drake TR-22, Motorola HT-220, and some Standard sets, I felt that the greatest need for more power would be from the one-watt level to approximately 30 to 35 watts, a power gain of approximately 15 dB. However, some readers are interested in somewhat higher input power ranges, so I have included basic designs for 1-watt in, 10-watts out, and 10-watts in, 35-watts out. Design and construction details, and tune-up procedures are the same as the 1-watt in, 35-watts out amplifier discussed in detail in the text.

Undoubtedly, more amateurs would attempt to build their own power amplifier if they could be assured of success upon completion of the project. With this in mind, I have included several design features that will greatly assist in realizing this goal, including the use of printed circuit inductances, a built-in vswr and relative power indicator to aid in tune-up and operation, and the use of state-of-the-art, high efficiency, rugged, yet low cost transistors. The manufacturer rates these transistors for *infinite* vswr, so you shouldn't burn them out if your antenna isn't what it's supposed to be.

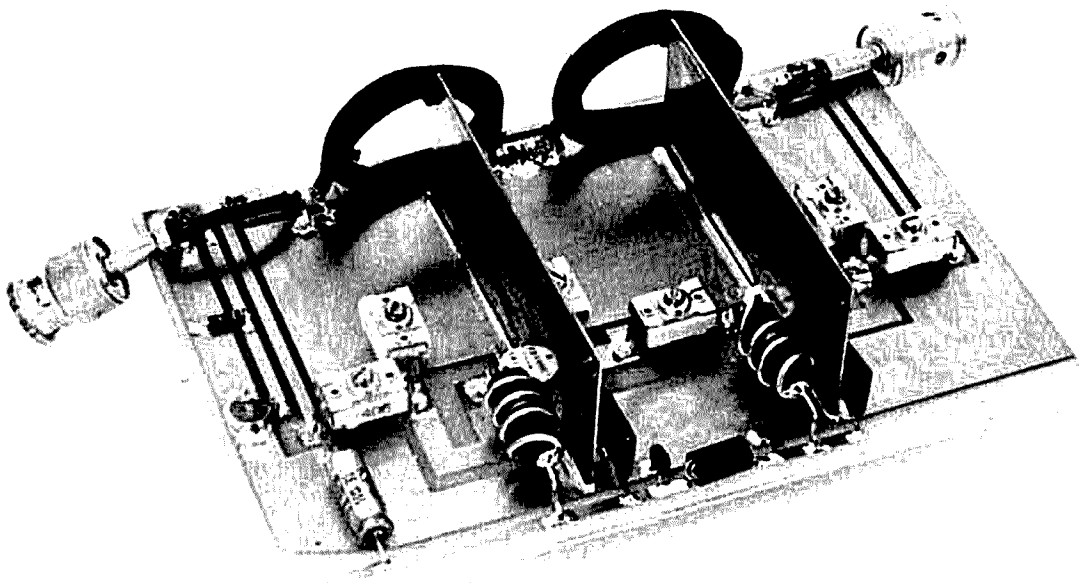
Although it will not be necessary for the builder to do any design work on these amplifiers, I will provide a short dissertation on the basic needs and techniques in dealing with high-power vhf transistor circuit design. I will also go through the complete procedure for designing the 1-watt input/35-watt output amplifier.

35-watt power amplifier

This amplifier is a two-stage solid-state unit with 30 to 35 watts output in the 146- to 148-MHz range when driven by 1 to 1.5 watts input. The 1-watt input is amplified to approximately 10 watts by the first stage, a SD-1198 transistor. This stage drives the final, a SD-1197 transistor, to approximately 35 watts output (depending upon how hot the transistors are). The output power passes through a vswr/relative power indicator printed on the board similar to the units described in a previous article.[1] From there it passes through two switching diodes to the output jack. The switching diodes are part of an automatic TR switching network developed by WB4-JMD and WB4DRB, and described previously.[2]

The theory of operation of this switching network is as follows: In the receive mode all six diodes are non-conducting and, as such, present high resistance blockage paths. When an incoming signal appears at the output of the amplifier, CR5 and CR6 block the received signal from entering the output circuit of the amplifier. The signal travels through the two quarter wave RG-174/U coaxial lines to the input jack of the amplifier. Since CR3 and CR4 are non-conducting they do not affect the passing signal. Since CR1 and CR2 are likewise non-conducting, the signal is thus prevented from entering the input stage of the amplifier.

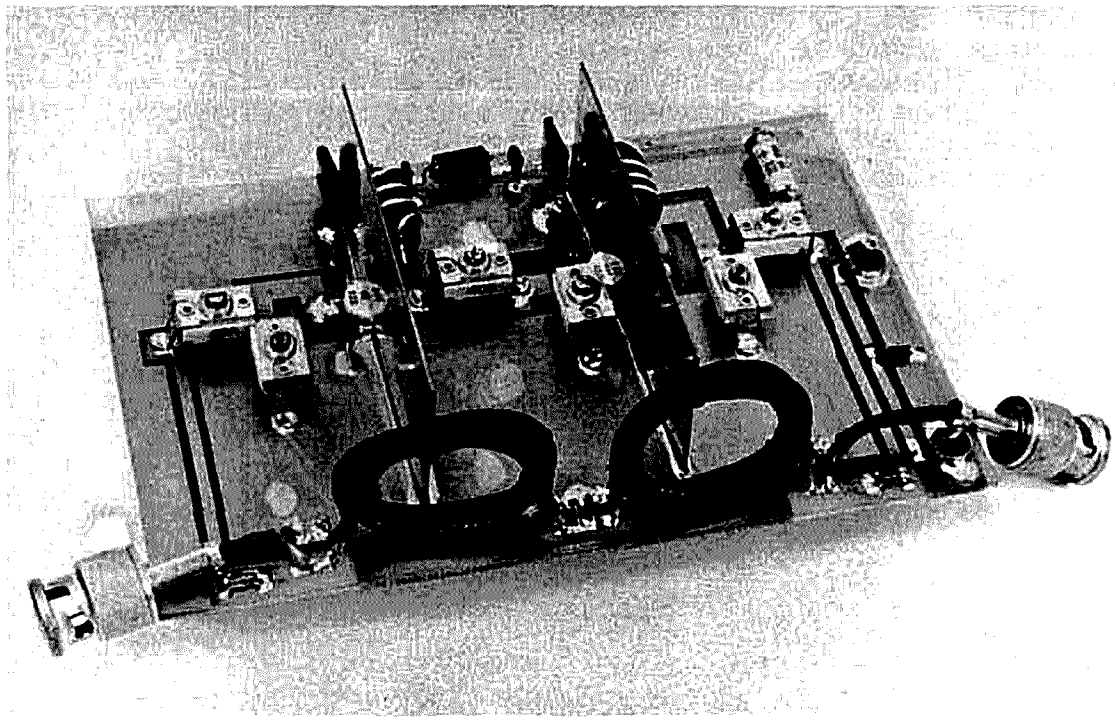
In the transmit mode, all six diodes are in conduction. Diodes CR3 and CR4 present a near short circuit to ground at the junction point of the two quarter-wave lines. Since a shorted quarter-wave line on one end looks like an open circuit on the other end, the line presents a very high impedance to the incoming driving



The solid-state two-meter power amplifier printed-circuit board includes etched inductors for both stages as well as a built-in swr bridge (upper left). Transmit-receive switching is accomplished with the two coiled quarter-wave lines at the rear of the board.

power. Diodes CR1 and CR2 are now biased into conduction and pass the driving power to the amplifier input stage relatively unobstructed. CR5 and CR6 pass the amplified power from the output stage to the output jack. Again, CR3 and CR4 perform the same function as in the input circuit and short out the output quarter-wave line, blocking the output power *signal*.

Since the amplifier is zero biased and displays excellent immunity to spurious oscillations, there is no need to switch the 13.6 Vdc supply voltage each time the driving transceiver is keyed. This fact, along with the solid-state switching network, yields an rf power amplifier free of any switching relays (often a source of trouble, particularly at these frequencies).



Built-in transmit-receive switch uses diodes and quarter-wave coaxial lines (foreground). In this view the input stage is on the left, the output stage on the right. The built-in swr pickup line is etched on the right-hand side of the board.

design considerations

Quite normally with any new design cost versus performance is one of the first considerations. It now appears that the Solid State Scientific Company's SD-1198 ($8.45) and SD-1197 ($12.30) vhf power transistors offer a good combination of reasonable cost and rugged construction while maintaining high performance in the power range of one watt in, 30 watts out.

After the determination has been made to use a certain type of transistor, the next step is to decide on what type of matching networks will work best. To do this, the designer must know the large-signal input and output impedances of the transistors he has chosen at the intended frequencies and operating power levels. Most manufacturers of high power rf transistors publish large-signal input and output matching information. Some companies omit the large-signal output real-part values that the transistor should be matched to. For these cases the designer may use the approximate equation:

$$R_{out} = \frac{(V_{cc})^2}{2P_o} \qquad (1)$$

Where V_{cc} is the dc input voltage, P_o is the rf power delivered to the load and R_{out} is the parallel output impedance in

ohms. For reasons that will be explained later, this parallel value, along with the output capacitance of the transistor, may have to be mathematically transformed into an equivalent series value. This can be accomplished by the following equations:

$$R_s = \frac{R_p}{1+(R_p/X_p)^2} \quad (2)$$

$$X_s = R_s \frac{R_p}{X_p} \quad (3)$$

Where R_s is the series resistance, X_s is the series reactance, R_p is the parallel resistance and X_p is the parallel reactance. Note that X_p is the transistor output capacitance in ohms at the frequency of operation.

Input matching information included on the manufacturer's specification sheet will include values for both resistance and reactance. Here again, if parallel values are given, it may be necessary, in some cases, to convert to equivalent series values.

Referring to the manufacturer's data sheets for the two transistors used in this amplifier, the input and output impedances at various power levels are given in series values:

SD-1198 input impedance at 1-watt input = 0.8 + j1.0

SD-1198 output impedance at 10-watts out = 4.0 - j1.2

SD-1197 input impedance at 10-watts input = 0.9 + j1.1

SD-1197 output impedance at 35-watts out = 2.2 - j0.1

matching networks

When the input and output impedances have been determined, the next step is to select the best type of matching network for the input, interstage and output. The two most commonly used networks are the pi and the tee. Theoretically, either network may be used, but the tee network will produce more reasonable values of inductance and capacitance when you're working with low values of transistor impedance. After selecting a value of circuit Q between 5 and 10, the L and C reactances for the tee network (see fig. 1) are calculated from the following formulas:

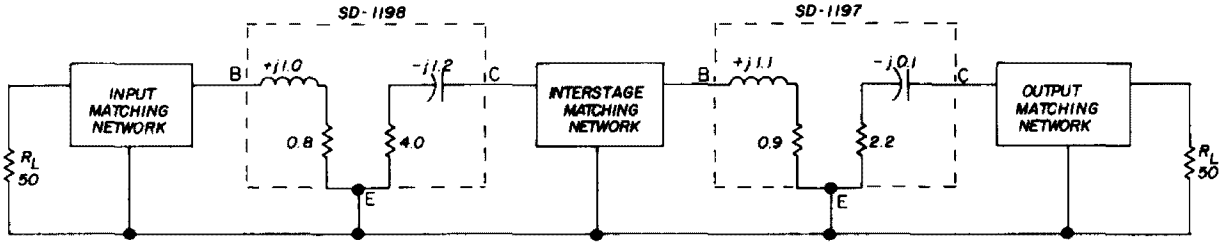


fig. 2. Series input and output impedances of the two transistors used in the two-meter power amplifier.

$$X_L = QR1 - [X1] \quad (4)$$

$$X_{C2} = AR2 + [X2]$$

$$X_{C1} = \frac{B}{Q \cdot A}$$

Where:

$$A = \sqrt{\left[\frac{R1\ (1+Q^2)}{R2}\right] - 1}$$

$$B = R1\ (1 + Q^2)$$

When you use these equations to

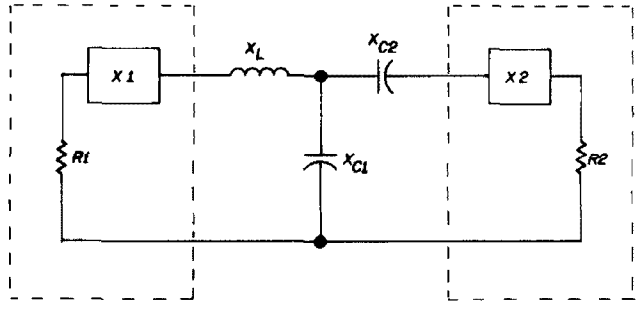


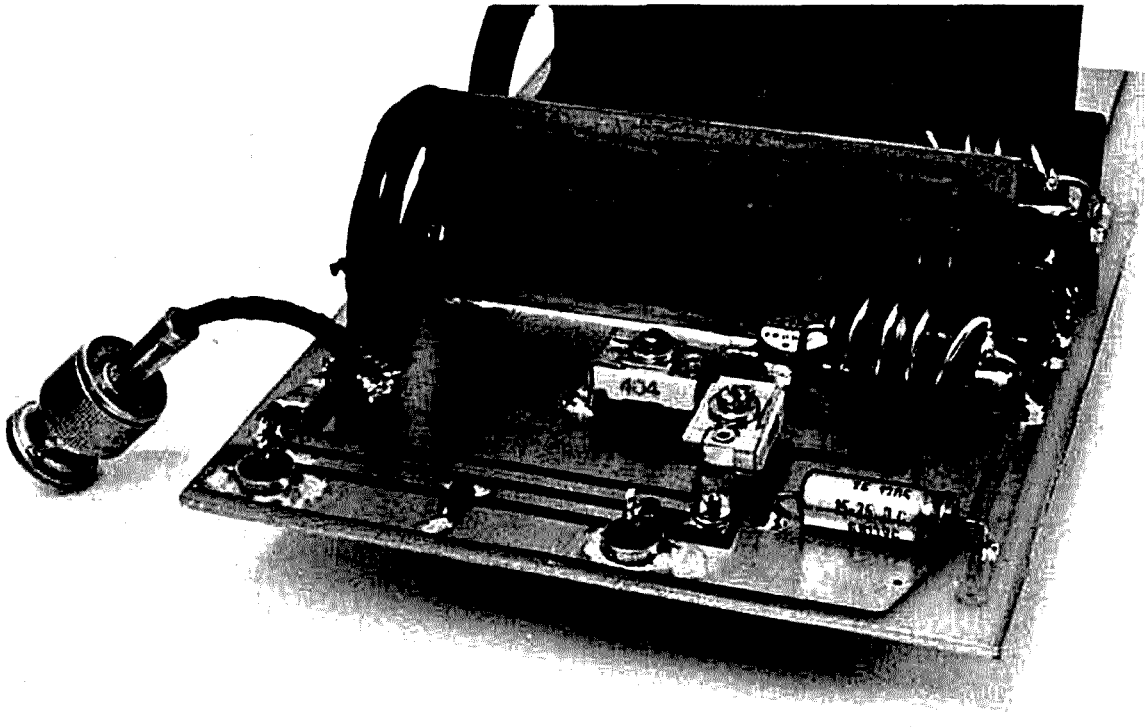
fig. 1. Tee matching network with series input and output impedances. Design formulas for this network are given in the text.

design a tee network, there are several rules which must be observed:

1. R1, X1, R2 and X2 must be expressed in series form. If they are given in parallel form on the data sheet, they must be converted to series equivalents using eqs. 2 and 3.

value except to stay within the 5 to 10 boundary. The designer arbitrarily picks the value he wants. It doesn't have to be calculated for this type of use. I picked a loaded Q of 10 for the input and output circuits, and a loaded Q of 7 for the interstage matching network.

The circuit for the input matching



Output tank circuit and swr bridge pickup line.

2. R1 must be lower in value than R2, and X_L must "look" into the lower of the two resistances. Arrange the tee circuit until this condition is met.

3. Enter X1 and X2 as a positive number if inductive; enter X1 and X2 as a negative number if capacitive.

network is shown in fig. 3. Since R1, the real part of the SD-1198 transistor series input impedance, is smaller than R2, the 50-ohm input impedance, the configuration of fig. 3 must be used for this matching network. Reactance X1 is the imaginary part of the transistor input impedance. Since there is no reactance at the input (see fig. 2) X2 is zero.

Using the values of fig. 3 and the formulas of eq. 4, you will arrive at the

At this point a word is in order about circuit Q. The Q in this case means the loaded Q, not unloaded. Experience has shown that a loaded Q between 5 and 10 works best in power amplifiers of this bandwidth (2 MHz at two meters). Too low Q will result in poor harmonic reduction and broad tuning. Too high Q will produce critical tuning, narrow bandwidth and poor transfer efficiency because of the high currents involved. Again, there is nothing magical about this

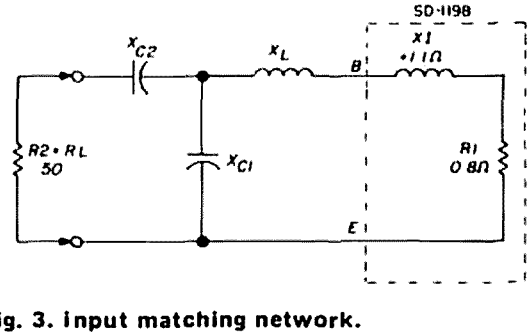


fig. 3. input matching network.

following reactance values for the tee network components:

$$X_L = 7 \text{ ohms}$$

$$X_{C2} = 39.24 \text{ ohms}$$

$$X_{C1} = 8.76 \text{ ohms}$$

The design of the interstage matching network shown in fig. 4 follows the same procedure. Using the values of fig. 4 in eq. 4 will yield the following values for the interstage matching network:

$$X_L = 5.2 \text{ ohms}$$

$$X_{C2} = 11.6 \text{ ohms}$$

$$X_{C1} = 11.8 \text{ ohms}$$

The same technique is used to design the output matching network shown in fig. 5. The reactance of the matching network components, from eq. 4, are:

$$X_L = 22.1 \text{ ohms}$$

$$X_{C2} = 92.8 \text{ ohms}$$

$$X_{C1} = 27.3 \text{ ohms}$$

The only remaining calculation is to convert all of the inductive and capacitive reactance values into inductance (in nanohenries) and capacitance (in picofarads). This is accomplished with eqs. 5 and 6.

$$C = \frac{1}{2\pi f X_C} \quad (5)$$

$$L = \frac{X_L}{2\pi f} \quad (6)$$

where C is in microfarads, L is in microhenries, f is in megahertz and π is 3.1416. Remember that there are 10^6 picofarads in one *microfarad*, and 10^3 nanohenries in one microhenry.

The complete two-meter power amplifier with matching networks is shown in fig. 6. Note, however, that the required inductance values are very small. Therefore, printed-circuit inductances are one of the most practical ways to obtain the required values. The graph in fig. 7 shows

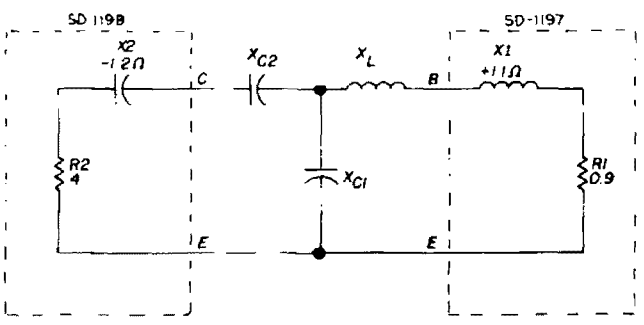


fig. 4. Interstage matching network.

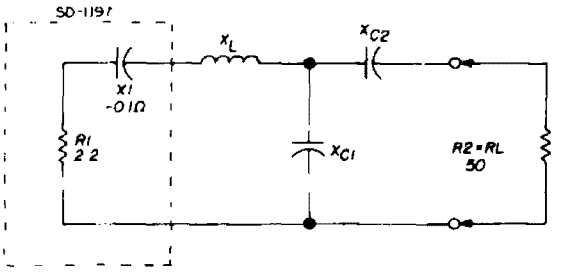


fig. 5. Output matching network.

the inductance of 0.1- and 0.2-inch copper strips on G-10 epoxy printed-circuit boards. The required length does not have to be laid out in a straight line, but can be etched in a loop as was done in the completed amplifier shown in fig. 8. When the printed inductances are etched as loops, the total length is measured on the center line of the copper strip.

Although the inductance and capacitance values shown in fig. 6 will be accurate in an ideal amplifier where all stray inductances and capacitances are eliminated, such will *not* be the case in a practical amplifier because, alas, we do not live in an ideal world. Because of the inherent strays, even in the best designs, the actual inductance and capacitance values will vary somewhat. That is why you must use variable capacitors in all three matching networks.

parasitics

When designing an rf power amplifier you must minimize the possibility of low- or high-frequency parasitic oscillations. Parasitic and spurious oscillations can be controlled by observing several important rules of thumb:

1. The impedance of the base-emitter choke should be no more than 50 times, and no less than 5 times the large-signal input impedance of the transistor.

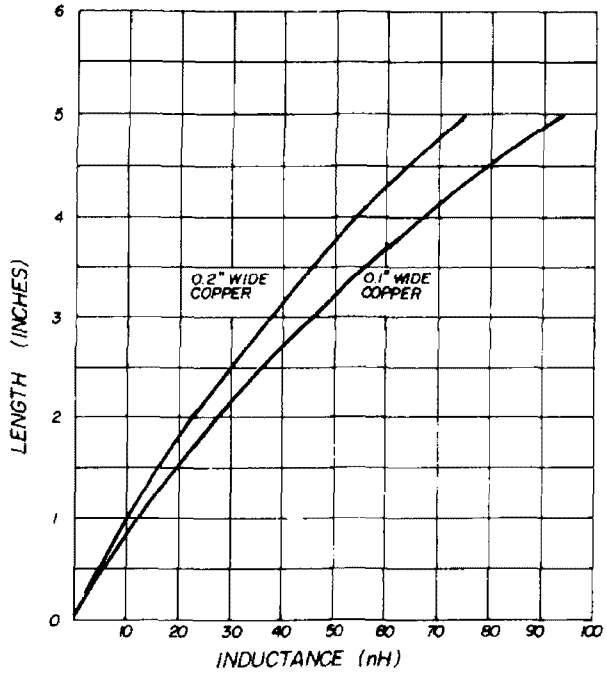


fig. 7. Inductance of 0.1- and 0.2-inch copper strips versus length. This graph may be used to determine the length of etched inductors on printed-circuit board (1-ounce copper).

2. The impedance of the rf choke in the collector circuit should be no greater than *10 times the large-signal* output impedance of the transistor (10 times the load impedance seen by the transistor at the fundamental frequency or the lowest frequency of operation).

3. It is a good idea *not* to use a ferrite rf choke in the dc collector circuit due to the possibility of the ferrite saturating and turning the stage into a blocking oscillator.

4. Use three separate bypass capacitors on the B+ supply lead consisting of 10-μF, 1-μF and 0.1-μF capacitors. The 0.1-μF capacitor should be as small as possible with the shortest possible lead lengths.

If you are designing a two or more stage power amplifier, a husky ferrite rf choke should be installed between the stages in the power supply line with adequate bypassing to ground. In extremely tough cases of spurious oscillation, a carbon resistor from base to ground may help. Its value should be no more than 50 times, or no less than 5 times, the large-signal input impedance of the transistor.

The use of a metal shield which isolates the input and output circuits will sometimes increase power output by reducing degenerative feedback. Finally, when you are designing an rf power amplifier, try to use the newest transistor types available. They may cost a bit more, but the results will be well worth it.

heat sinks

A good, conservative rule of thumb for proper heat sinking is to use a heat sink capable of dissipating as much or more power than the amplifier produces. Silicon heat-sink grease should be used between the transistor header and the heat sink to insure maximum power transfer.

Use caution when tightening a transistor to its heat sink because the metal used in the threaded studs of most rf power transistors is very soft and can be easily twisted off if you use too much torque.

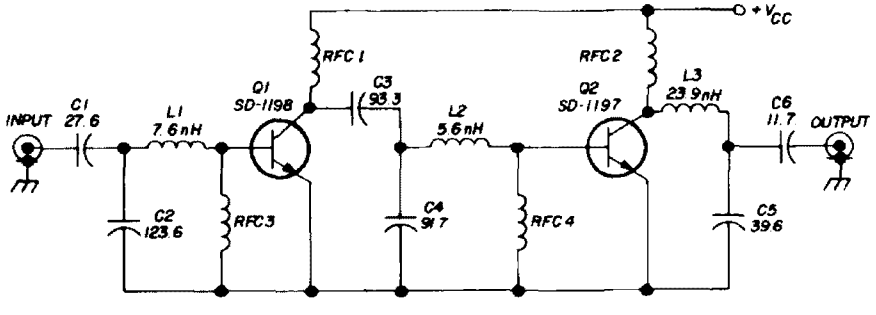
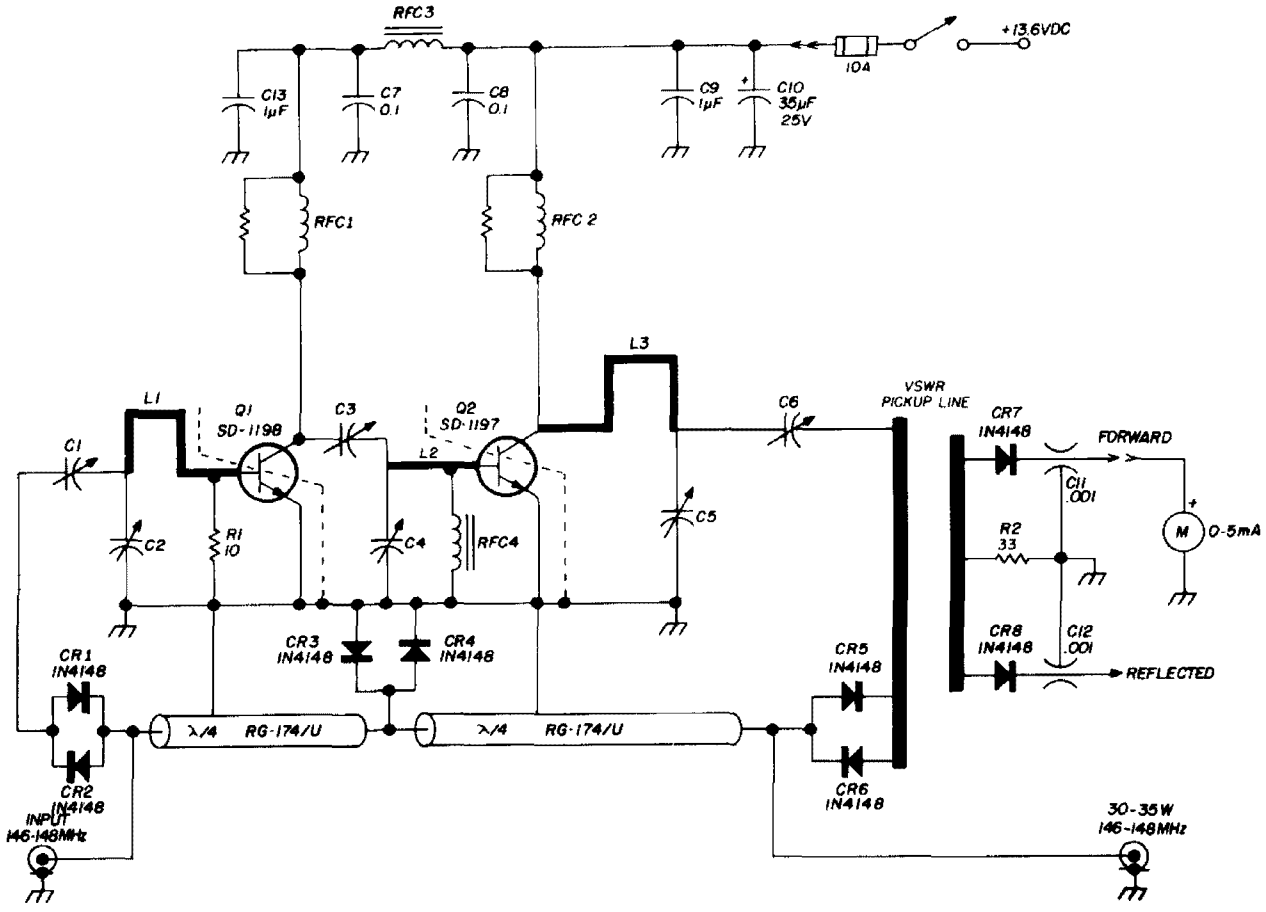


fig. 6. Basic design of the two-stage rf power amplifier.

construction

The two-meter rf power amplifier shown in fig. 8 is built on a G-10 epoxy



C1,C3	15 to 115-pF trimmer (Arco 406)	CR1-CR8	1N4148 (Texas Instruments)
C4,C6		L1	7.6 nH printed inductor
C2,C5	4 to 60-pF trimmer (Arco 404)	L2	5.6 nH printed inductor
C7,C8	0.1 μF, 50 V (Erie 8121-050-651-104M)	L3	23.9 nH printed inductor
C9,C13	1 μF, 50 V (Erie 8131-050-651-105M)	Q1	SD-1198
C10	35 μF, 25 V electrolytic (Sprague TE-1208)	Q2	SD-1197
C11,12	.001 μF feedthrough (Allen-Bradley FA5C-102W)	R1	10 ohms, ¼ watt, 5% carbon
		R2	33 ohms, ¼ watt, 5% carbon
		RFC1,RFC2	4 turns no. 18 on 470-ohm, 2-watt, carbon resistor
		RFC3,RFC4	ferrite choke (Ferroxcube VK-200-10/3B

fig. 8. Complete two-meter power amplifier with etched inductors, etched vswr/power indicator and solid-state TR switching. The two quarter-wave RG-174/U coaxial lines are each 14-3/16" long.

printed-circuit board with printed inductances and a printed vswr and relative power indicator.* The vswr and relative power indicator is similar to one previously described.[1] The use of a printed-circuit board allows the builder to produce a reliable power amplifier as close to the original design as possible with a minimum potential error.

The shield partition between the stages is made from 0.010-inch copper or brass, and is laid out as shown in **fig. 9.** The

*Completely etched and drilled printed-circuit boards for these amplifiers may be obtained from Artronics Company, Box 462, Merrifield, Virginia 22116. The board for the 1-watt input/35-watt output amplifier is priced at $13.50; boards for the other two units are priced at $11.50, each, postpaid.

The power transistors and other components for these amplifiers are available from Almo Electronics, Roosevelt Boulevard and Bluegrass Road, Philadelphia, Pennsylvania 19114. The price for the SD-1198, 10-watt power transistor is $8.45; the SD-1197, 35-watt power transistor is priced at $12.30.

heat sink is a 6-inch section of Wakefield 640 heat-sink extrusion, or equivalent.

Start construction by trimming ail transistor leads to 3/16-inch long. Lightly tin the area of the PC board where the transistor leads and variable capacitors

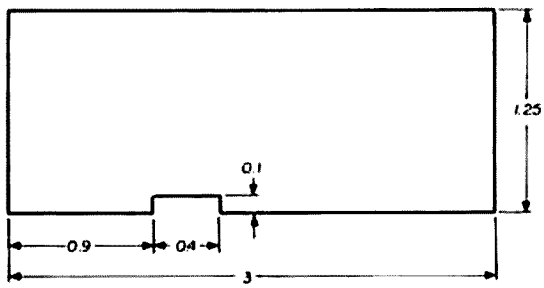


fig. 9. Layout of the copper or brass shield partitions used between stages (two required).

will be installed. Bend all variable-capacitor leads at the mid-point of the cutout slot. This will keep the adjustment screw from shorting out to ground after the capacitor is installed.

Install both transistors and solder them securely in place. Sweat solder the two shield partitions in place, and sweat solder the two .001-μF feedthrough capacitors in place and cut off the protruding lug on the heat-sink side of the PC board.

Now install the rf chokes, base resistors, bypass capacitors and switching diodes, including the two used in the vswr indicator. Install the 33-ohm resistor in the vswr circuit. Install the six variable capacitors and adjust C1 and C6 to maximum capacitance. Install the two quarter-wave RG-174/U coaxial lines.

Attach the printed-circuit assembly to the heat sink using Dow silicon heat-sink grease (or similar). Use caution when you tighten up the transistor stud nuts, as mentioned previously.

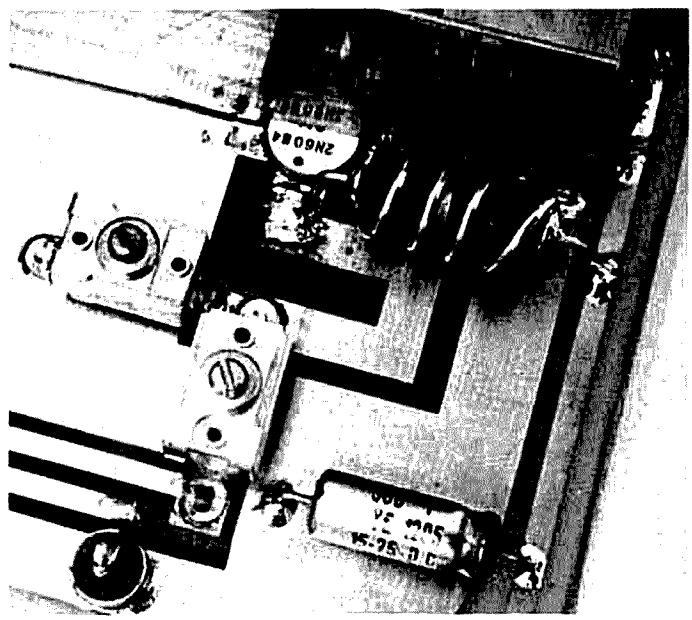
tuneup

Temporarily connect two short coaxial cable leads (RG-174/U) to the input and output points of the amplifier. Connect the output lead to a good non-reactive 50-ohm dummy load such as a Bird Termaline®, or perhaps, a Heath Cantenna. Do not use a light bulb as it is too reactive at this frequency.

Connect the input to a suitable 146-148-MHz driver with 1 to 2 watts output. Connect the B+ lead through a 7-amp quick-blow fuse to preferably a variable power supply. Connect a volt-ohmmeter (5 mA range) between the vswr feedthrough capacitor nearest the output transistor and ground. Slowly raise the supply voltage to approximately 10 to 11 volts. If there is no indication on the meter, a more sensitive range may help.

During initial tuneup it may be easier to monitor the dc ammeter on the power supply and tune for maximum current, up to about 4 amperes. By that time there should be an indication on the rf power output meter. From this point on all adjustments should be for maximum indication on the rf output meter. A reading of approximately 4 mA will indicate maximum rf power output of 30 to 35 watts. This should pull 5 or 6 amperes from the dc power supply.

The completed rf power amplifier can be installed in a Minibox or similar enclosure with the heat sink installed on the outside. Two test jacks may be brought out of the cabinet for monitoring vswr and output power. The 13.6-volt B+

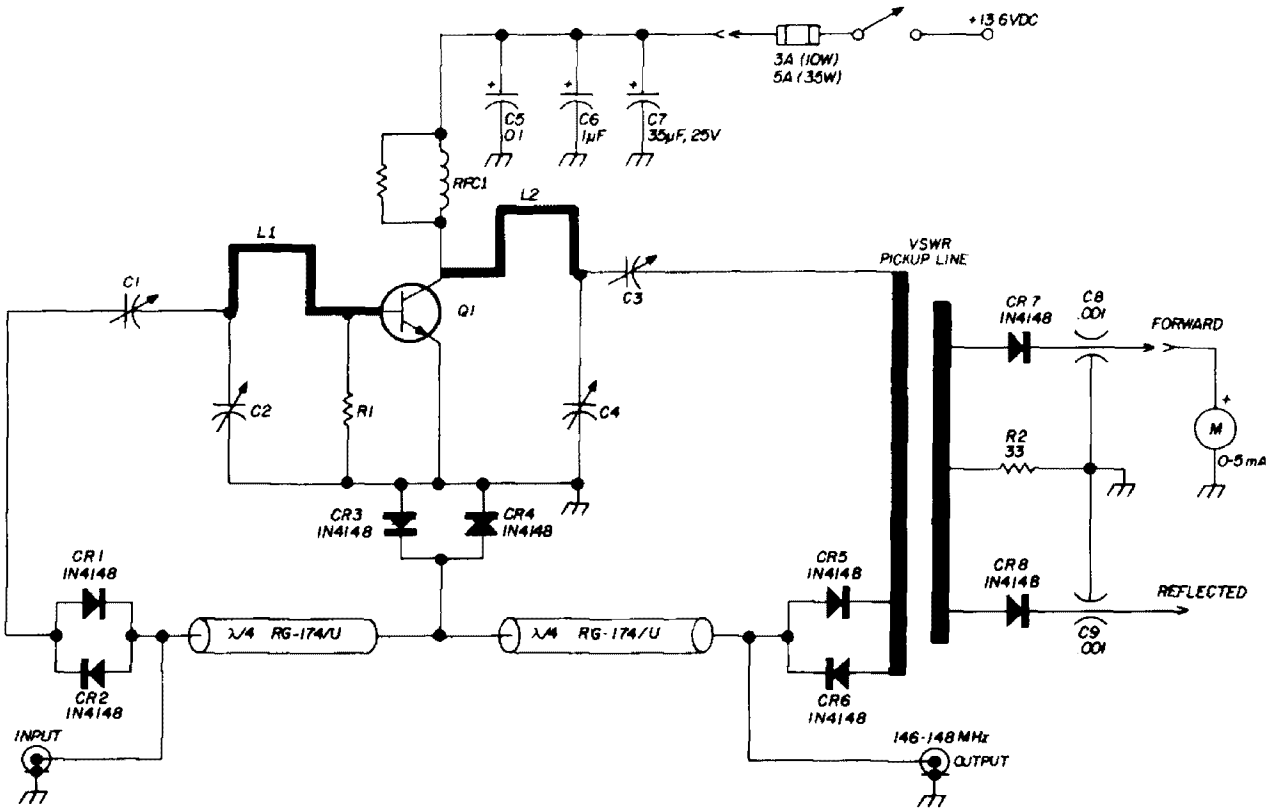


Closeup of 35-watt output stage shows construction simplicity that results from using printed-circuit inductors.

lead should be in series with a switch and a 10-amp quick-blow fuse to your automobile battery or other power supply.

acknowledgements

I would like to thank Chris Galfo, WB4JMD, and Marc Dressman, WB4DRB, for permission to use their solid-state TR switching technique. I would also like to thank John Gregory, W3ATE, for providing the photo-copy work for the printed-circuit board.



C1,C3	15 to 115-pF trimmer (Arco 406)
C2,C4	4 to 60-pF trimmer (Arco 404)
C5	0.1 μF, 50 V (Erie 8121-050-651-104M)
C6	1 μF, 50 V (Erie 8131-050-651-105M)
C7	35 μF, 25 V electrolytic (Sprague TE-1208)
C8,C9	.001 μF feedthrough (Allen-Bradley FA5C-102W)
CR1-CR8	1N4148 (Texas Instruments)
L1	10-watts output, 7-nH printed inductor 35-watts output, 8-nH printed inductor
L2	10-watts output, 28 nH printed inductor 35-watts output, 15 nH printed inductor
Q1	SD-1198 for 10-watts output, SD-1197 for 35-watts output
R1	10 ohms, 5% carbon. 1/4-watt for 10-watts output, 1/2-watt for 35-watts output
R2	33 ohms, 1/4 watt, 5% carbon
RFC1	4 turns no. 18 on 470-ohm, 2-watt carbon resistor

fig. 10. Single-stage rf power amplifier for 1-watt input, 10-watts output or 10-watts input, 35 watts output. The two quarter-wave RG-174/U coaxial lines are each 14-3/16" long.

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ham radio

the vertical radiator

A discussion of the operating characteristics of simple vertical antennas for 40, 80 and 160

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On the 40-, 80- and 160-meter amateur bands, a vertical radiator is superior to the horizontally-oriented dipole. It gives omni-directional coverage with extremely low radiation angles when anchored in the earth, which will provide excellent long-range communication and good signal strengths at short range.[1] Further, if properly constructed, it will provide full band, low vswr coverage over the three amateur bands.

balanced horizontal radiator

The balanced horizontal radiator system is one of the most common systems in use. If a half-wave dipole antenna is horizontally oriented in space, free of all other objects, the radiation pattern is like a doughnut at right angles to the axis of the dipole wire. However, if the dipole is situated near the earth (quarter wavelength or less), the ground reflected ray cancels most of the direct ray of radiation at lower vertical angles. Only the higher angles are propagated with appreciable amplitude under these circumstances.

If the dipole is oriented about a half wavelength above the earth, the radiation pattern has two fairly broad major lobes at right angles to the wire in the vicinity of 25-35° above the horizontal. To reduce the major lobe to 10-20° the dipole has to be elevated to approximately one wavelength above the ground. Dipoles have little radiation below 15° in any direction if less than one wavelength above poor earth.

A typical half-wave dipole for 80 meters would be about 125 feet long. If it were made of no. 10 wire (0.1-inch diameter) it would have an extremely high characteristic impedance (over 2000 ohms). This would make an efficient radiator at resonance; however, due to its high characteristic impedance, it would have a very narrow operating bandwidth.

If the two halves of the dipole could be made into a multiwire cage about one foot in diameter, the characteristic impedance would be approximately halved and the bandwidth increased proportionally.

unbalanced vertical radiator

If, instead of a balanced dipole, an unbalanced system fed by unbalanced coaxial cable were used, and further, if the outer conductors of the coaxial cable were unbraided into a plane at right angles to the center conductor, an unbalanced radiator would result. This form of radiator is known as a monopole or unipole. If the center conductor is vertically anchored in the earth it becomes known as a vertical monopole, for obvious reasons. If the length of the center conductor is made an electrical quarter-wavelength long and the plane of outer conductors is also made one-quarter wavelength, it becomes a very efficient omnidirectional radiator at low vertical angles.

When the center conductor is about one-quarter wavelength, the vertical pattern is quite broad and extends from near 10° to about 50° at the half power points with the peak radiation at about 25° above the horizontal. When the length approaches 1/2 wave, the vertical pattern is concentrated between 30° and near 5°, with the peak of radiation near 15° above the horizontal. At 5/8-wavelength, the lobe is concentrated between zero and 20° with the peak radiation about 8° above the horizontal. Additionally, at this height there is a small high-angle lobe near 60°.

At lengths in excess of 5/8 wavelength the low-angle lobe diminishes in amplitude and the high-angle lobe increases until, at a height of one wavelength, the low-angle lobe disappears altogether and the high-angle lobe is at a maximum. See figs. 1 and 2 for a comparison of the vertical pattern of a typical vertical radiator and a horizontal dipole.

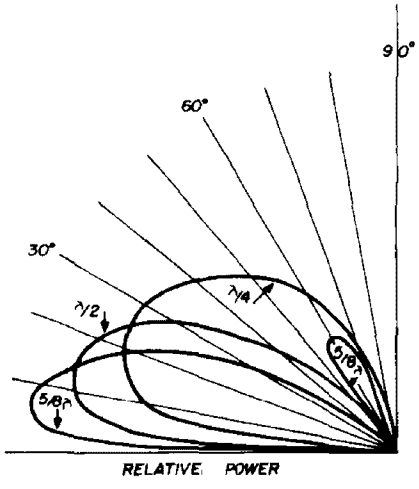


fig. 1. Radiation pattern of the vertical monopole antenna. The pattern is omnidirectional for heights of 1/4, 1/2 and 5/8 wavelength when an adequate ground plane is used.

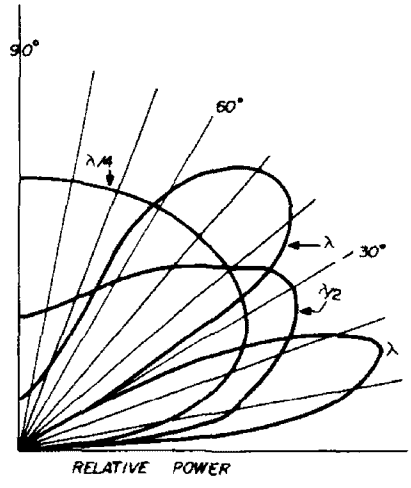


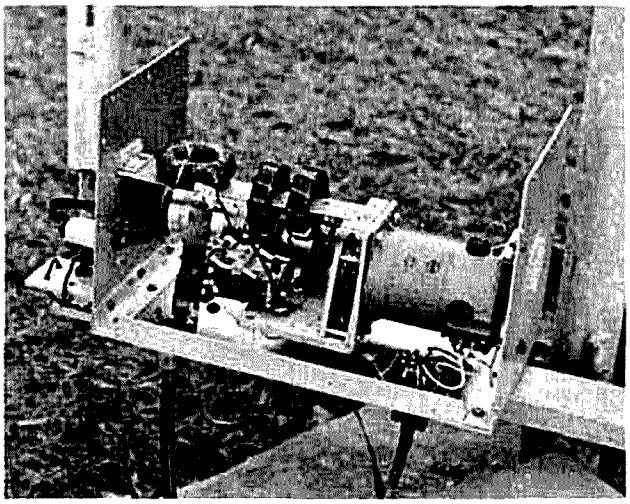
fig. 2. Radiation pattern of the horizontal dipole antenna. Pattern at right angles to the wire for heights of 1/4, 1/2 and 1 wavelength over a good ground.

If the vertical (center) conductor is small in diameter with respect to its length, its characteristic impedance would be a high value. Typically, a quarter-wave vertical conductor at 80 meters, made of 63 feet of no. 10 wire, would exhibit a characteristic impedance of over 1000 ohms. On the other hand, if the vertical conductor were to be made of a metal pole, 1 foot in diameter (30-inch diameter triangular tubing tower is the equivalent), it would have a characteristic

impedance near 500 ohms. (This is about one-fourth the characteristic impedance of a thin dipole.) This would result in a much broader bandwidth characteristic and allow the use of the radiator throughout an amateur band without appreciable variation in its electrical characteristics. The bandwidth available for efficient operation is proportional to the radiating conductor's diameter to length radio.

ground plane

Other factors which affect the efficient operation of a radiator are the ohmic (rf) resistance of the system and dielectric losses induced by material of lossy characteristics between the two halves of the radiator. No one in his right mind would erect a dipole and then surround one of the two elements with several inches of lossy material such as grass, soil or clay! Ridiculous? Not at all! For years the designers of vertical radiators have been burying the outer conductor extension of the unbalanced vertical system in the earth "to get a good ground."



Control box located at bottom of the tower. The geared motor on the right drives the voltage-variable capacitor which is nearly hidden by the two toroid inductors. The circuit for the control box is shown in fig. 3.

A good *ground* we do not want. A large plane of highly conductive material that the vertical radiating member can see without intervening lossy dielectric *is* what is needed! The current must flow between the two halves of the system without lossy material in the intervening space, and charge the medium that will carry the signal outward. The *ideal* outer conductor extension would be a continuous plane, at least one-quarter wave long of metal of high conductivity that would completely hide the earth in the near field (most model studies are made this way). Since this approach results in rather poor economics at the lower amateur frequencies a mat of mesh wire and/or a radial wire system, extending outward one-quarter wavelength in as many directions as the budget and available space will allow, will provide an adequate outer conductor extension of the unbalanced system. However, the mesh/wire should *not* be buried deeply.

Since the typical amateur's backyard location of an antenna system must be available for other purposes, as a compromise the radials may be placed into the sod of the grass, as shallow as is possible. They should not present an obstacle course, nor should they be buried so deeply that part of that expensive antenna power is used in warming the back yard.

How extensive should the outer-conductor extension plane be made? First, let's review the things not to do. It should *not* be made by merely connecting the bottom of the vertical member to several rods driven into the earth to get a good ground. Five ground rods 6-feet long (or three ground rods 10-feet long) connected to the tower base is good lightning insurance, however.

The extension plane need not be made one-quarter wavelength long in all 360° directions of the radial system. If an omnidirectional pattern is a requirement it should be about $\lambda/4$ in all directions but this is not mandatory if some pattern distortion is permissible or desirable. The individual wires need not be of large diameter if there are enough of them to satisfy the area and conductivity requirements indicated below.

Now, let's think of some of the things that should be done. The outer conductor extension place of the unbalanced vertical

radiator should be made of material that has high conductivity. Also, between the base of the vertical and the extension plane the connecting conductors should be made of many large cross-section wires to reduce the chance of losses in this high-current part of the system.

The extension plane should have an area (total) approximately equal to the area of the vertical member. To build a capacitor with several square units of area on one plate and a single square unit of area on the other would not be good practice. A radiator is an L-C circuit; the length of its elements provide the inductance and the area of its members provide the capacity. Therefore, they should be roughly equal in area and length. If necessary, use wire matting (fence wire) near the tower base (10 to 15 feet out) to augment the radial wires.

If the radial wires are spaced not more than 0.1 wavelength at a distance of $\lambda/4$, they will intercept about 95% of the energy (16 radials meets this requirement). This is not a perfect design but will provide a good extension plane for the outer conductor. If the mesh mat is made of fence wire of the type that has each crosswire welded to the longitudinal wires it will provide a high capacitance at the base of the vertical radiator.

If some directivity is desired it may be concentrated in one direction both in length and spacing. Do not overdo the length, however. There is a point of diminishing returns beyond 3/8-wavelengths long.[1]

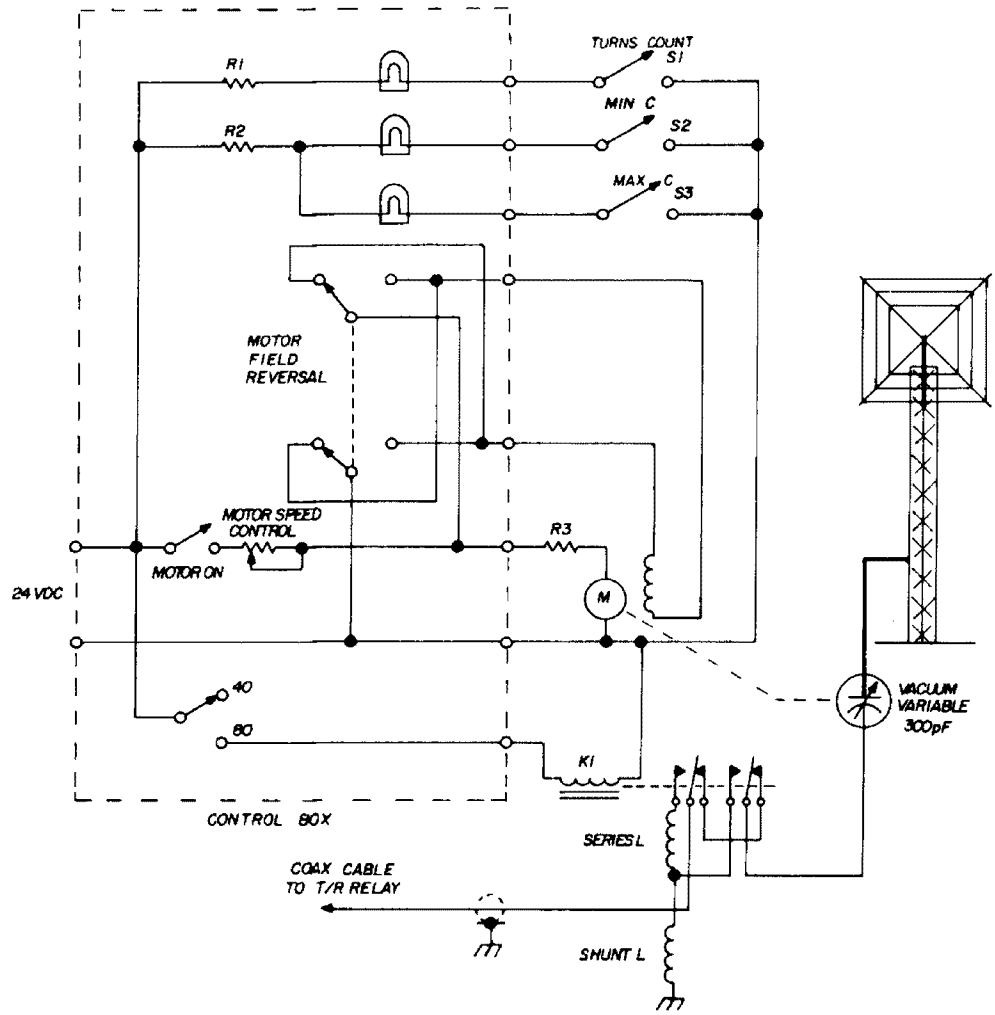
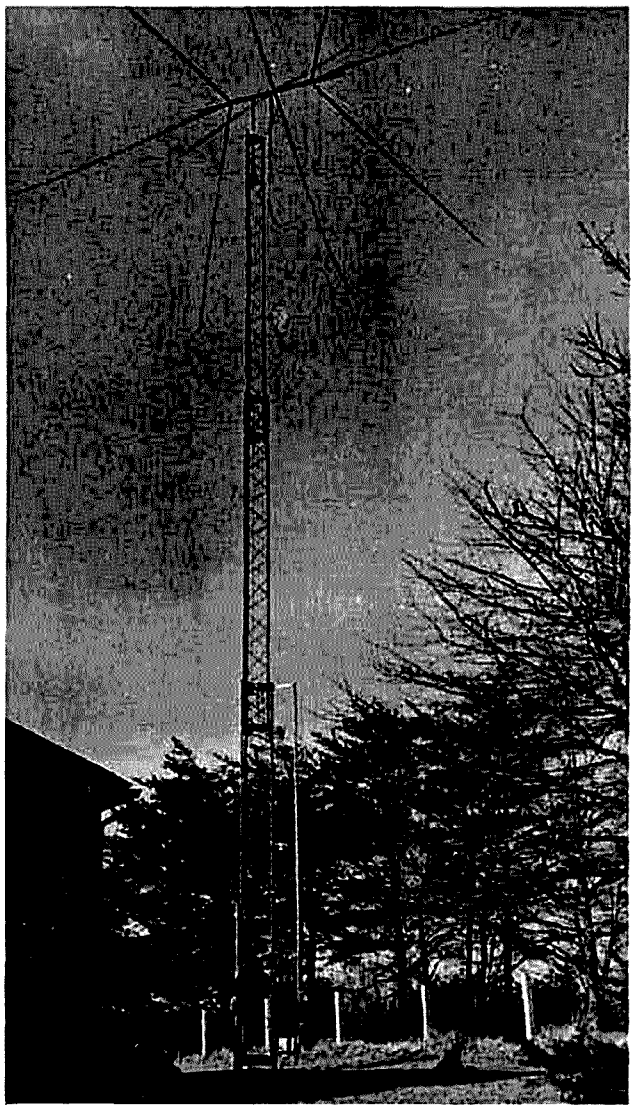


fig. 3. Electrical control system for the 40- and 80-meter vertical antenna. Microswitches S1, S2 and S3 are limit switches actuated by the motor system, which control the pilot lamps. R1 and R2 are pilot-lamp current-limiting resistors. R3 controls maximum motor speed. The motor is a 24-volt unit geared down to drive the vacuum variable capacitor.

Excellent results can be obtained by using multielement, multiband, rotatable beams on 20 meters and above, especially if they are mounted on top of a tower that gets them far enough above the earth for low angle radiation (50 to 60 feet). They can be purchased or constructed at a nominal price. Also, they can be made rugged enough to withstand rather heavy weather.



The W40Q vertical-tower antenna system that covers all bands, 80 through 10 meters.

For the lower frequencies it is nearly impossible to get a large enough beam, with rugged characteristics, up a full wavelength in height for low-angle radiation, without excessive expenditure of funds and effort.

These are the reasons that this article is directed toward the use of the vertical radiator on 40 meters and lower. Further, the use of a beam supporting tower for the vertical member of the radiator is recommended as a bonus with minimum additional expenditure.

base impedance

There are many charts which show the characteristic impedance, as well as the resistive and reactive components, for the *ungrounded* vertical radiator when fed at the base. However, there are several reasons why the ungrounded vertical radiator is impractical for use by the radio amateur. For one thing, any static level and/or lightning strokes in the vicinity would channel this collected energy directly to the equipment to which it is connected. Also, any use of the tower for other purposes, such as for mounting hf or vhf beams, would require all the cables to be routed through a decoupling device at the base so that these cables would not short out the base insulators. The tower would almost certainly have to be guyed in order that too much stress will not be placed on the base insulators.

Whether the tower is ungrounded or grounded, towers that require guys should be insulated from metallic guys. These guys must be broken up with egg insulators into non-resonant lengths to reduce pattern distortion. Non-conductive guys of polypropalene are relatively invisible to rf at high frequencies and are reputed to be quite strong and inexpensive. Nylon stretches too much to be considered for use in tower guying.

Also, whether the tower is guyed or unguyed, any rf and control cables from the beams down to the base, should be routed inside the tower structure all the way to the bottom to reduce pattern distortion. Additionally, they should be routed underground a few inches from the tower base to the operating position. Here is a good place to use that lossy earth. It will attenuate any rf on the outside of the cables. This will also assist in TVI reduction.

There are few charts available for feed impedance versus tower heights for the grounded tower vertical radiator. Al-

though there are too many variables in such a system to make an accurate prediction of the feed impedance some generalities can be stated.

A grounded-base tower can be easily shunt fed by the use of either the delta-or the gamma-type feed system. The delta system will radiate a little and distort the radiation pattern. Therefore, it is not considered practical. The gamma system is an excellent choice for shunt feeding the grounded radiator. Its feed impedance can be adjusted by changing the length of the vertical member. It can also be adjusted a small amount by changing the diameter and spacing of the gamma rod. In all cases where the gamma rod is less than one-quarter wavelength long, it will show inductive reactance. This reactance can be cancelled by a series variable capacitor. Typical R ± jX readings for various lengths of gamma rod and vertical antenna heights are given in table 1.

table 1. Typical resistance and reactance values for various lengths of gamma rods and vertical antennas.

tower height (wavelengths)	gamma length (wavelengths)	resistance (ohms)	reactance (+j ohms)
0.25	0.1	5-10	200-300
0.5	0.2	40-80	700-1000
0.5	0.25	over 500	over 1000'

note: The above readings were made with a fence-wire ground mat (two lengths, 100-feet long by 3-feet wide). The tower height was varied from 22 to 54 feet (a 3-element tri-band quad mounted on top). Another tower, ground plane or beam setup will have a different feed-point impedance but it should be near those figures listed above.

From table 1 it becomes obvious that a gamma length too near one-quarter wavelength is to be avoided, especially if the vertical radiator is near one-half wave. At the other extreme, if the gamma rod is much shorter than 0.1 wavelength and the tower is well below one-quarter wave high, the feed-point resistive component may be so low that transformation to the coaxial-line impedance may be difficult.

If all three lower frequencies available to the amateur are to be used on one vertical antenna, it is recommended that the two higher frequencies be used with one gamma rod, and that a second gamma system be provided on the opposite (physical) side of the tower to reduce reaction between them. This second gamma system can be made physically longer on the tower than the other one which will increase the resistive component.

measurements and matching

After the gamma matching system has been installed the measurement of feed point reactance and resistive components can be accomplished by use of a commercial or home-made rf impedance bridge.[2] Once the electrical characteristics on each amateur band have been determined, the design of the reactance cancellation and resistance matching transformation to the coaxial feed impedance can be solved by the use of a graphical technique.[3] This technique requires no more exotic tools than a straight edge, a compass and graph paper. The mathematics required will not exceed simple arithmatic. Every amateur interested in antenna work should add this valuable set of tools to his experience and knowledge.

Next month I will present some practical solutions to the construction of the vertical radiator and its ground plane. The use of the rf impedance bridge and solution of impedance matching by use of the graphical method will also be shown. A practical control system using a motor-driven vacuum-variable capacitor to cancel the reactance will be explained. Additionally, a switching system that will allow complete remote control of one vertical radiator over the 40- and 80-meter amateur bands will be illustrated.

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ham radio

phasing-type ssb generator

Complete construction information on a simple, all-passive phasing-type single-sideband generator

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Phasing-type single-sideband generators described in the literature use active devices, either to isolate the phase-shift networks from the modulators or as the modulators themselves. A few years ago it occurred to me that it should be possible to build a phasing-type sideband generator using only resistors, capacitors, inductors and diodes and using no dc power. This article describes a recent version of such a unit.

Its input requirements are about 3 volts peak rf across 50 ohms (0.1 watt) and a few tenths of a volt of audio across about 15 ohms. Output is from an rf transformer whose tuned winding can drive a vacuum tube or fet directly, while a transistor can be driven from an added low-impedance winding or a capacitive tap on the tuned winding. An attractive aspect of the circuit is that it could be fabricated as a compact, general purpose, sideband generator three-port, since only the two rf phase-shift reactors and the output transformer depend upon carrier frequency.

Usually, both phase-shift networks are designed for infinite impedance loads, automatically requiring that tubes or fets follow. On one memorable occasion, however, a design appeared for a relatively low-impedance network where the input impedances of the transistor loads were included in the resistive output branches of the audio phase shift network.[1] My unit carries this idea to the logical conclusion of dispensing with the drivers and simply including the impedances of the diode modulators in the design and adjustment of the output branch of the audio phase-shift network. The rf phase-shift network design also allows for the modulator impedance. Finally, since the diode modulator impedances vary somewhat with rf drive level, some swamping is used to simplify design and adjustment.

The result appears in **fig. 1.** The version shown operates at 10.5 MHz. Other versions have run at 12 and 7.25 MHz. The values shown for the audio phase-shift network components are those suggested by van Heddegem. The values actually used were scaled up a few percent in impedance to accomodate a couple of odd precision resistors in my own junk box. The test equipment used to build and adjust this circuit consisted of a diode rf probe connected to an ordinary vom, a low-frequency R-C bridge[2] and a receiver with a mechanical filter to verify sideband suppression.

The method of sideband switching is one I have not seen mentioned in print. It has the advantage that one side of every modulator input may remain permanently grounded. The only audio transformer required is the one driving the phase-shift network.

circuit description

The circuit consists of three sections, the audio and rf phase-shift networks and

the two balanced modulators. The rf phase-shift network values are those I used, but other builders should be guided by what identical precision resistors are available. The diode balanced modulators are so arranged that both rf and audio inputs are single-ended with one side grounded. The audio and rf inputs are effectively in series, as usual.

The rf phase-shift network is an ordinary arrangement of plus and minus 45-degree sections, except that an estimate of the resistance looking into each modulator is made part of the resistive branches. To estimate this resistance, note that each diode conducts half the time. With one diode conducting, the rf input resistance for each modulator is half the balancing pot, 500 ohms, plus a few hundred more ohms for the diode (varying somewhat with rf drive level) plus another 500 to 1000 ohms for the load seen at the output winding feeding the rf transformer. This load will depend on the Q of the rf transformer and on the load presented by the other balanced modulator. You can only guess at these values, but a reasonable guess is that they will nearly match the roughly 700 ohms of diode plus balancing pot, giving an estimate of about 1400 ohms.

The 150-pF bypass capacitors were made small to avoid affecting the performance of the audio phase shift network. They have about 100 ohms reactance at 10.5 Mhz but do not significantly affect the rf phase shift, since they are in series with a few thousand ohms as well as combining in quadrature. Furthermore, any effects they have are identical for the two modulators, shifting phase slightly without changing the 90-degree relationship.

If 51 ohm resistors are used in the phase-shift networks, the modulator resistance is swamped by a factor of about 28, so a 28% error in the estimate of the modulator resistance is equivalent to a 1% error in the rf phase shift resistor. For this combination the effective resistances of the rf phase shifter output branches are about 49.2 ohms; the rf phase shifter L and C should have this reactance at the operating frequency.

My network was built by making the C from a precision unit a little above the calculated value in series with a larger low-precision unit. The L, a slug-tuned coil, was then adjusted so the rf voltages across the two phase shifter outputs were equal. As a check, note also that each of these outputs is about 70.7 percent of the whole rf input voltage. There is no reason why both L and C should not be variable. In that case you can forget about the exact modulator impedances as long as

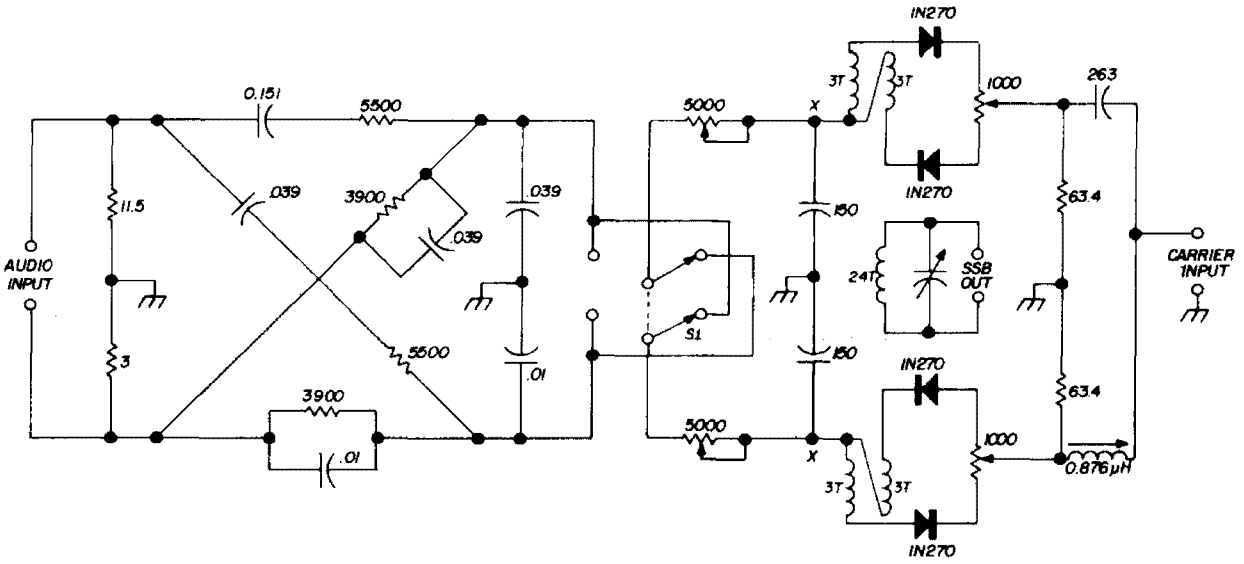


fig. 1. Circuit of a passive phasing-type ssb generator which uses no active devices.

the two are built identically, but the rf phase-shift adjustment becomes more complicated.

audio phase-shift network

The audio phase-shift network was made of junk-box items. Resistors were precision types or a precision type shunted by a much larger ordinary resistor. Capacitors were paralleled from whatever was on hand. All R and C values were checked on the bridge before being wired in. I've built three audio phase-shift networks by this *dead-reckoning* method (no adjustments) and they all seemed to work fine as judged by listening to speech while switching sidebands.

The resistances at the input of the audio phase shift network should have a ratio of 3.83 with 1% accuracy, but actual values don't matter. I used 5.24 and 20 ohms because my junk box supplied them. Note that the sum of these two resistances should be matched to the audio driver. A junk output transformer from a midget broadcast set is ideal, since it has about the right turns ratio to match a medium-mu triode to 15 or 30 ohms and operating it at this higher impedance level will reduce low-frequency response below 300 or 400 Hz. Be sure neither old vc lead is grounded when using such a transformer.

The important adjustment in this unit is to make each output resistive branch look like the desired 3900 ohms. Each of these resistances is the sum of the trimpot resistance and the modulator audio input resistance. The modulator resistances will be about 700 ohms since the modulator rf transformer windings have negligible audio impedance. This shows that there is a swamping factor of about 5 for variations in the audio impedance of a modulator.

To insure that the resistance of this branch is correct under operating conditions, disconnect the input end (end at the dpdt switch) of each trimpot from the audio phase-shift network and connect this end of the trimpot and ground to the R-C measuring bridge. Set the bridge dial to 3900 ohms (or whatever the design resistance is for your own audio phase-shift network). Turn on the rf carrier oscillator to be used with this unit. It should be operating at the intended voltage, preferably regulated (I'm using a 6C4 fed from a VR150). Now adjust the trimpot for a bridge null. Repeat for the other trimpot and both output branches will have the desired effective resistance.

To insure proper combination of the two modulator outputs, all coils of the output transformer should be closely coupled and both sets of modulator windings should be identical bifilar windings. The easy way to do this, used in the unit described, is to wind the output rf transformer on a high-permeability toroid. The bifilar modulator windings are made from a slightly twisted pair of a few inches of thin magnet wire (cotton or silk covering increases your confidence in the absence of shorts) closely wound around the core.

The two modulator windings and the tuned output winding are all slightly spaced from one another so coupling will be mainly from flux through the core. The numbers of turns indicated in fig. 1 worked at 10.5 MHz for a nice mystery core from the junk box. For your own core an easy solution is to use a grid-dip meter and the expected tuning capacitance (add 10 pF or so for strays) to discover how many turns are needed on the tuned winding. The modulator windings are then chosen for the indicated turns ratio.

Another arrangement I have used successfully is a solenoidal tuned winding with the two modulators connected to a quadrifilar winding made from four wires slightly twisted together and wound over the solenoid near the middle. A slug-tuned form is convenient in this case. The turns ratios are not critical. Cup cores are nice if you have small, steady fingers.

Balance pot adjustment also deserves some comment. Originally the usual 1000-ohm pots shown in the diagram were used. Later these were replaced by 100-ohm pots at the centers of pairs of 470-ohm resistors to ease carrier nulling.

This required some diode selection, unnecessary with the 1000-ohm pots. The diodes are 1N270s removed from computer boards, but any rf germanium diode is suitable.

In this diode balanced-modulator circuit an rf bypass is often shown across each 1000-ohm balance pot. Such bypasses will cause balance to occur far from pot center unless the diode back resistances are matched or swamped by about 50k across each diode. The reason is that this rf bypass assumes a steady voltage that applies reverse bias to the diodes. Since the diodes are in series for this bias, they divide it in proportion to their back resistances. Unless they are well balanced or swamped to equality the result is to require an eccentric setting of the carrier balance pots. The rf portions of the unit should be shielded so that balance is unaffected by hand capacitance.

sideband selection

The usual method of switching sidebands in a phasing-type sideband generator is to reverse the phase of one of the audio signals or one of the rf carrier inputs. This is not practical when all modulator inputs are to be single-ended with one side at ground. A dpdt switch could be used to reverse the connections of one pair of diodes to the corresponding modulator output winding, but this destroys the symmetry of the circuit and might degrade sideband suppression. At this point in thought I noticed that interchanging either the two audio signals or the two rf signals would also switch sidebands. Obviously, the audio signals are the ones to switch. I have never seen this method of sideband switching in print but assume the experts are aware of it.

results and afterthoughts

When I built the present unit, the rf phase shift C was computed and a fixed capacitor wired in; the coil was then adjusted to make the two modulator rf input voltages equal. The audio phase-shift network trimpots were then adjusted as described earlier and a voice-modulated signal tuned in on a receiver with a 3-kHz wide mechanical filter. The signal sounded normal. When the sideband switch was flipped only a few faint glitches were audible, indicating satisfactory sideband suppression. As noted, the test equipment used on this project is too rudimentary to permit any fancier assessment of unwanted sideband suppression. Suppression was audibly the same on either sideband position.

Having finished the unit and gotten it working, I noticed that I'd forgotten to include an audio balance adjustment to equalize the contributions from the two balanced modulators. In general such an adjustment should be included. Don't try to do it by twiddling the trim pots in the audio phase shift network. A suggested method is about 10k from one of the points X to ground and 5k pot plus 7.5k resistor from the other point X to ground. These resistors should be in the circuit with the variable branch set at 10k when the trim pots are adjusted.

The very small adjustment that is made to reach exact audio balance will not significantly affect the operation of the audio phase-shift network. If audio balance can't be reached (poor unwanted sideband suppression) look elsewhere for the trouble. A test of the audio phase-shift network can be made by checking that the output voltage at each trimpot (end away from modulator) is exactly half the voltage across the 3-ohm input resistor.

This unit has been on the air with conversion to the 75-meter band, where it was reported as sounding like a good, normal ssb signal. Of course, it also sounded normal, as monitored on my own receiver, before it was put on the air.

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ham radio

rf phase meter

An instrument for measuring the phase angle between elements of vertical antenna arrays

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Vertically polarized directional antennas are popular for use in the 40- and 80-meter amateur bands.[1] Successful operation of these antennas depends on the phase relationship between antenna elements. Described here is an instrument that permits phase-angle measurements within acceptable limits for aligning such antenna systems.

Computer studies[2] have shown that satisfactory antenna performance can be obtained if the lead-lag phase angles are within ±5 degrees of the target (design) figure and the current ratios are within ±5 percent of the target ratios. The phase meter described will perform within these limits to 7 MHz. The circuit uses standard components and is easily calibrated with a vtvm and oscilloscope. The circuit also has inherent lead-lag discrimination.[3]

The phase angle, $\emptyset$, by which waveform A leads B (**fig. 1**) can also be expressed as a time differential, Δt. The phase meter measures Δt, which can be read in terms of an average dc voltage at the output.

functional description

As an aid to understanding circuit functions, the truth tables and timing diagram of **figs. 3** and **4** should be used in conjunction with the block diagram, **fig. 2**.

Channels 1 and 2 are identical up to the Exclusive-OR gate. The input to the Schmitt trigger is provided by the pickup loops, which are not part of the circuit but are placed near the applicable antenna element. The Schmitt trigger squares the positive half of the incoming sine wave, and its output is fed to flip-flop 1.

A flip-flop is essentially an electronic

switch. The devices each have two output terminals, one of which is on while the other is off. Every negative-going input pulse will cause the flip-flops to switch state. If, for instance, output A is on, then the next negative-going input pulse will switch A off and B on. The next negative-going input pulse will then switch the device to its original state; i. e. A to on and B to off. Thus, if A is the output terminal, it will be on for every second input pulse, and the input frequency will have been divided by two in the output. Two flip-flops are used in series to divide the signal frequency by four. The first flip-flop in Channel 1 is used to provide gating pulses for each NAND gate.

The Exclusive-OR gate provides an output if either A′ or B′ is present, but not if both are present or both are absent. Therefore its output consists of pulses exactly as wide as interval Δt, or the lead-lag angle, $\emptyset$.

A NAND gate will always have an output except when both inputs are present. The timing diagram shows that for a Channel 1 input leading that of Channel 2, NAND gate 1 output will be at a constant value, which is that of the logic on value. This average dc level will be directly proportional to the lead angle of Channel 1; it will decrease with increasing lead lead angle. If, on the other hand, Channel 2 input leads that of Channel 1 NAND gate 2 output will be creasing lead angle. If, on the other hand, Channel 2 input leads that of Channel 1 continuously on, while the average dc level of NAND gate 1 will vary with lead angle. Thus, the circuit provides a linear measurement of the lead-lag angle between two waveforms and also allows an unambiguous determination of which waveform leads or lags in relation to the other.

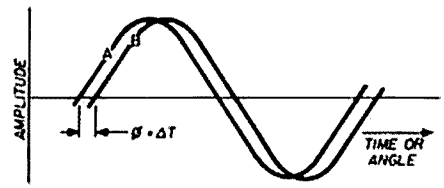


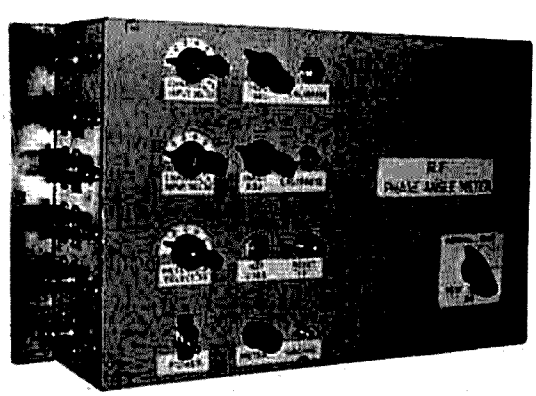
fig. 1. Relationship between two sine waves in terms of phase angle, $\emptyset$, and time, Δt.

circuit details

The phase-meter schematic is shown in fig. 5. A single pickup loop (for simplicity) is shown connected to the system. The pickup loop acts as a current transformer looking into coaxial lines terminated in their characteristic impedance, Z_O (R1-R6). Potentiometers R20, R21 are used to reduce and equalize input voltages to the logic circuits. The MC1550G and the 2N2369 operate as a combined Schmitt trigger (wave squarer) and interface to drive the following TTL stages.

The circuit was originally developed for use up to 5 MHz.[4] As modified here it's not a true Schmitt trigger but will produce reasonably square waves to 10 MHz. Calibration potentiometers R11, R12 are used to compensate for unequal phase shifts through each channel.

The frequency dividers and subsequent logic circuits must be connected as shown, otherwise errors and confusion in the phase lead-lag determination will result. The output metering points have RC filters that respond to the average value of the square waves. Various RC combinations were tried and did not seem critical. The circuit will work without the filters, but as a matter of principle it's not good practice to have rf hash floating around the chassis. Test switch S4 is used to check the operation of the Ex-



Top view of the assembled phase meter.

clusive-OR circuit, and reset switch S5 is used to synchronize the channels.

power supply

The power supply is important because the calibration limits of the phase detector (Exclusive-OR) circuit output are 0.09 and 3.8 volts. Originally an unregulated supply was used. During the calibration of the meter these limits changed with variation in the house-current supply.

Since the power supply is mounted externally, the protective diodes are needed in case of wiring errors. The various voltage-dropping resistors ensure equal positive and negative voltages on each of the triggers and give the approximate 3 volts needed for the HI inputs of the flip-flops and gates. If you wish to use a built-in ±5 volt regulated supply, then only R26 and R27 will be needed.

component notes and layout

Details of the pickup coils are shown in **fig. 6**. With the transmitter running at 40 watts input, sufficient voltage was available to drive the meter. Although the length of the cables between coils and meter is not important, *all* cables must be the same length.

It is absolutely essential that the terminating resistors, R1-R6, are non-inductive. We used the meter described in reference 5 to check a couple dozen carbon ¼-watt resistors to find six suitable 52-54 ohm components.

The meter has six inputs rather than three so that phase angles in the antenna power divider could be measured as well as those between the antenna elements. The meter has two types of input connectors, which further enhances its versatility.

Wiring to the input selector switches must be as shown to avoid phase-shift errors (see the schematic and the bottom view of the chassis). Diode probe circuits C1, CR1, R7 (channel 1) and C2, CR2, R8 (channel 2) must have leads as short as possible. However, the wiring between R7, R8 to switch S3 can be 4 inches long, unshielded.

As can be seen in the photos and the component layout sketch, **fig. 7**, most of the circuit is on a PC board fitted with twelve 8-pin, dual in-line sockets. The

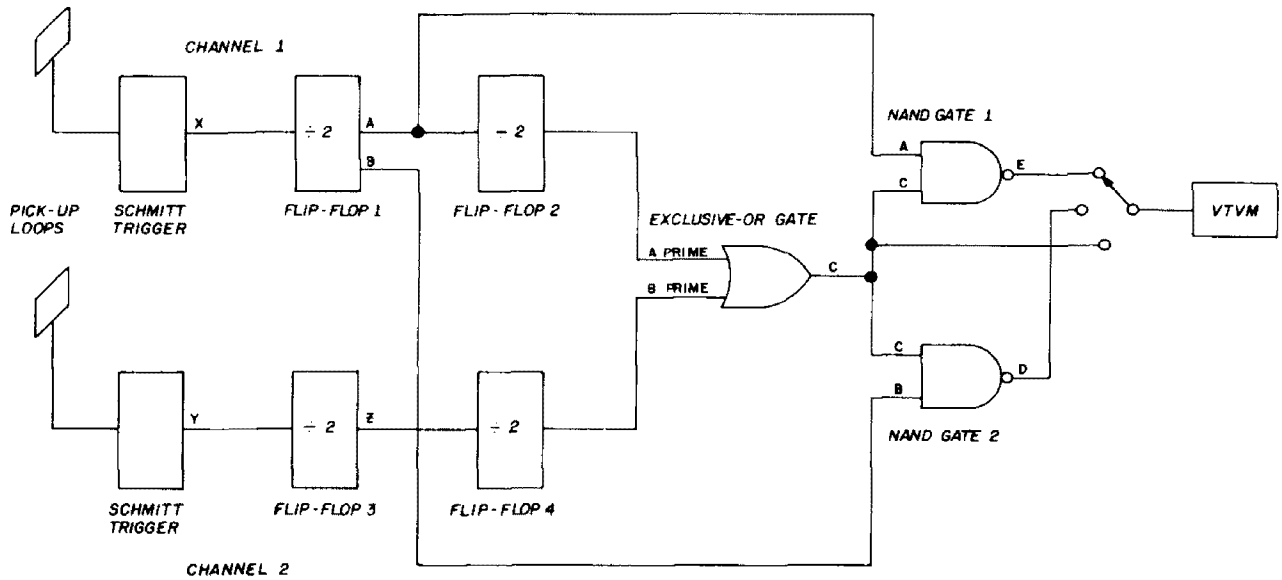


fig. 2. Functional block diagram.

Input A C	output E	Input A' B'	output C	A' B' C A B D E
0 0	1	0 0	0	0 0 0 1 0 1 1
0 1	1	0 0	1	0 0 0 0 0 0 0
1 0	1	1 0	1	1 0 1 1 0 1 0
1 1	0	1 1	0	1 1 0 0 1 1 1
NAND gate 1		**exclusive-OR**		**combined logic**

fig. 3. truth table.

board happened to be on hand and permitted about a dozen different variations in circuit layout, which accounts for the general disorder shown. If the project were to be started again, stick-on "Circuit Zaps" and a perforated board would be used. Note that wiring must be as direct as possible for the rf circuits. For example, switch S4 must be mounted near the flip-flop to which it's connected.

Voltage-dropping pots and isolating chokes are on a separate board and its location isn't critical. Number 26 hookup wire was used throughout, and coax was used only at the switches and Schmitt trigger inputs.

Carbon resistors rated ¼ watt should be used, because they perform well at high frequency. Pots R11, R12, R20, and R21 should be carbon. Chokes L1, L2 aren't critical but are needed to isolate the Schmitt triggers.

initial adjustments

The only test instruments needed to align the phase meter circuits are a scope with a 5-MHz bandwidth, an rf probe, a Heath vtvm, and an rf signal generator.

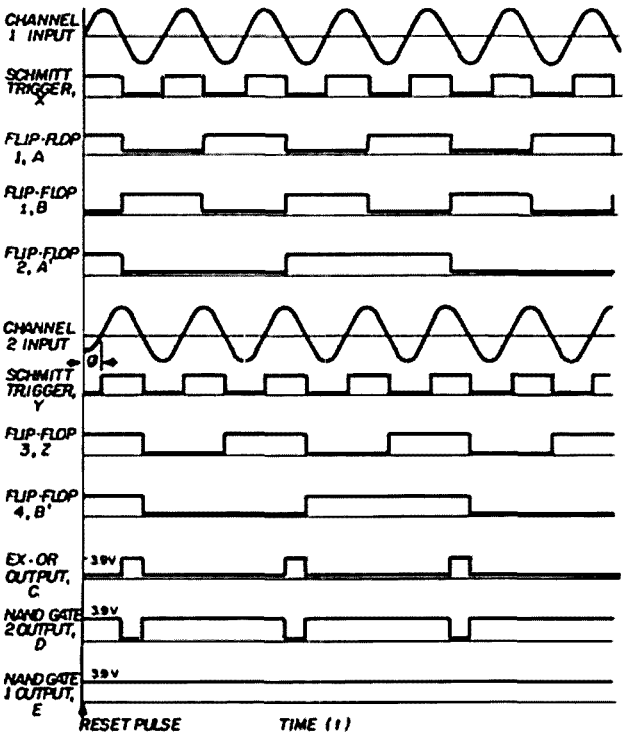
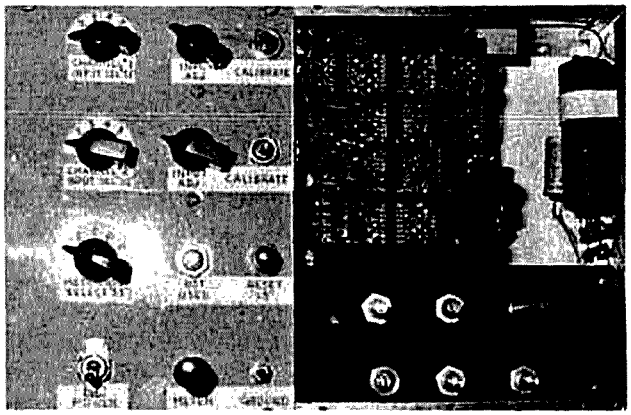


fig. 4. Timing diagram. In this example channel 2 lags channel 1 by Ø = 90 degrees. Note that the average values of C, D, E correspond to those in the calibration curve, fig. 8.

The following procedures should be followed for initial adjustments.

Before plugging in any devices, turn on the power and ensure that the correct polarities appear at the various device terminals. The power supply used here



Top view with cover removed.

provided ±5.8 volts at the input to the phase meter circuits. When all devices were installed, the voltage appearing after the protective diodes was about ±4.9 volts.

After checking voltages, turn off the power, connect the vtvm to the meter terminal of S3, and plug in the 7473s and the EX-OR gate devices. Adjust R26 for 3.0 volts at the EX-OR gate HI outputs. With S3 at position C and S4 in the normal position, the vtvm will indicate either 0.09 volt or 3.8 volts, which are the LO and HI outputs of the EX-OR gates. Pressing reset switch S5 will cause the 3.8 volts to decrease to 0.09 volt; or, if the reading is already 0.09 volt, pressing S5 will make no difference. Very often when the power is switched on, the reading will be 0.09 volt and no amount of pressing the reset switch will produce 3.8 volts; this is normal. With switch S3 still at position C and switch S4 in the opposite position, the meter will read 3.8 volts, and pressing the reset switch will have no effect.

Return S4 to normal and turn S3 to D and E, which are the NAND gate outputs. Normally 3.8 to 3.9 volts (HI) will be found, and no amount of pressing the

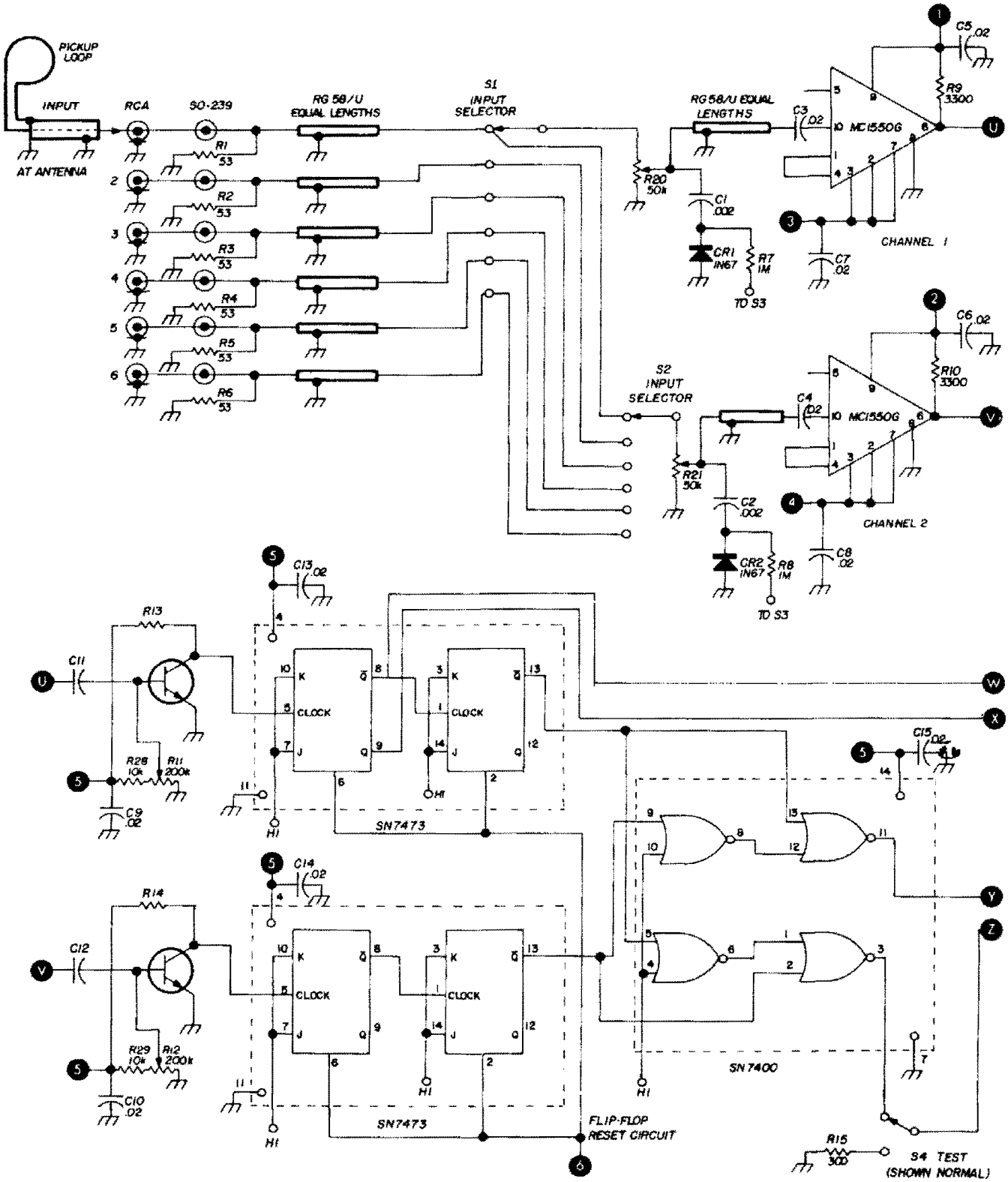
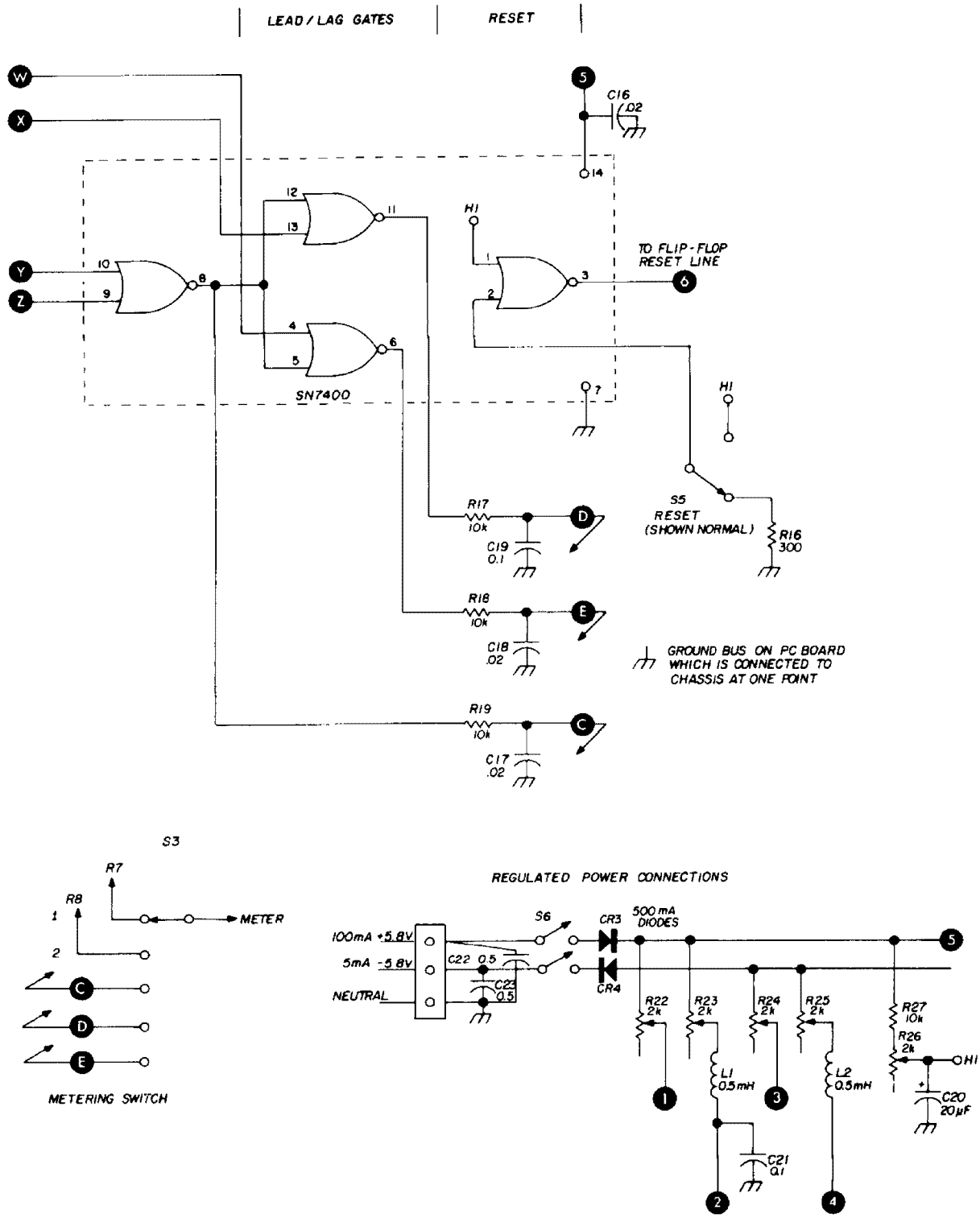


fig. 5. Phase meter schematic.

reset button should change this. With S3 at the opposite position, the meter will read the LO of the NAND gate. Note that the values of HI and LO depend on IC construction and the values mentioned here just happen to be what were found with these particular devices. If you don't get the above sequence of voltages you've made a mistake somewhere, and any further steps will be just a waste of time. Plug in the MC 1550s and 2N2369s and adjust the supply voltages by means of R22, 23, 24, 25 so that positive and negative voltages are equal at the IC terminals.

It's possible to adjust the instrument



at 7.5 MHz, but if difficulties are found it's easier to see where you're going at a lower frequency of, say, 1 MHz, where a relatively inexpensive scope will give a good picture. Hence the following procedure is based on 1 MHz, and one channel at a time will be aligned.

With S1 at Input 1 and R20 at maximum voltage, connect a signal level to produce about 0.2 Vdc on the meter. With the scope probe at the output of FF 2 (pin 13) you may be lucky and see some fuzz on the first shot. If not, slowly turn R11 one way or another until you do. Then increase the frequency slowly to 7.5 MHz and, if the fuzz disappears, readjust the pot until it reappears. Check that pressing the reset switch will make the fuzz disappear. If the fuzz cannot be controlled by the pots, then lower the

values of R28 and R29 slightly.

Repeat the procedure for Channel 2 with a signal at input 2 and S2 at Input 2. Next, put the two channels in parallel via S1 and S2, and feed in a signal at 1 MHz

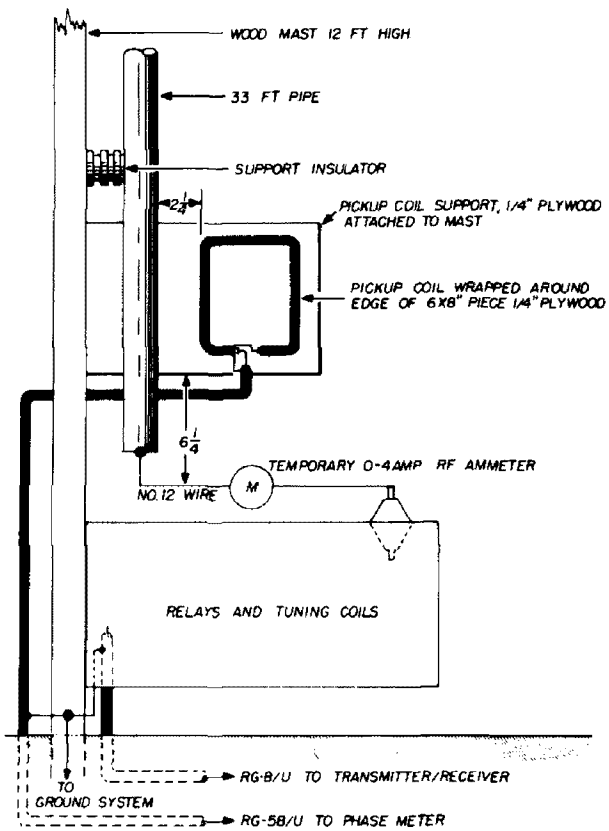


fig. 6. Antenna pickup loop details. The pickup coil must be in the plane formed by the L of the support pipe and the no. 12 wire.

(the pot setting for 7.5 MHz will also be good down to 1 MHz). Adjust R20 and R21 for equal voltages with switch S3 in positions 1 and 2 and switch S3 to C. The reading here will likely be anything up to 3.8 volts, and the object is to get it as close to 0.09 volt as possible. Press and release the reset button to ensure that 0.09 volt is possible. By carefully readjusting R11 and R12 the reading can be brought down to 0.09 volt. Then, when frequency is increased to 7.5 MHz, this reading will rise to about 0.18 volt. Further adjustment of R11 and R12 will bring this down to about 0.12 volt, which is at zero degrees of the calibration chart, fig. 8. Values of 3.7 and 3.9 volts should appear at positions D and E and these are also on the chart. If you can't get C down to 0.12 or 0.11 volt, then something may be wrong with metering circuits R20, C1, R7.

The meter is working in every respect when the above readings are obtained. Input signals as low as 0.05 volt and as high as 0.6 volt gave equal phase angle readings for a given length of test cable. A somewhat annoying feature of the basic circuit is that it covers 720 degrees, as shown on the timing diagram. This means that a phase angle of, say, 30 degrees gives a reading of 0.25 volt and also a reading of 390 degrees at about 2.05 volts. Hence, when the zero adjustment is made, the meter will occasionally pop up to 1.8 volts, and the reset switch must be pressed to bring it back to the proper value. This causes no ambiguity because the D and E readings change accordingly and, as will be explained in the next section, the sum of any two valid C readings must be around 1.88 volts.

calibration

The secret is to find six readings that correspond to a known lag or lead condition and plot the data. For example, with a signal frequency of 3.8 MHz fed to Channel 2 and a cable length of 45 ft. 9 in. connected between Channel 1 and Channel 2, Channel 2 will lead Channel 1 by 96.5 degrees, assuming a cable velocity factor of 0.66.

table 1. Channel 1 and 2 phase angle as a function of calibration voltage.

test	switch 3 position	meter reading (volts dc)	
	C	0.55	
1	D	3.34	Channel 2 leads Channel 1 by 96.5 degrees
	E	3.39	
		1.33	
		1.88*	
2	D	2.28	Channel 2 lags Channel 1 by 263.5 degrees
	E	4.03	

*Sum of any two C readings (validity check)

After adjusting the Schmitt trigger inputs to equal values (0.26 volt) and momentarily pressing the reset switch, the readings in test 1, **table 1**, were noted. Pressing the reset switch four or five times changed the values to those shown at test 2, **table 1**.

From the timing diagram and the magnitude of the HI and LO voltages of the EX-OR gate, it is known that a 90-degree lead will correspond roughly to 0.5 volt at C, 3.3 volts at D, and about 3.9 volts at E. The discrepancy at E in test 1 (**table 1**) is annoying but poses no difficulty. The values found during test 2 were about the expected complementary values. Note that the sum of the C readings is always about 1.88, and this fact is used as a validity check at all times.

The six values are plotted on the calibration chart, **fig. 8**. Pieces of coax of 4, 11 and 23 feet were on hand and were combined in various ways to obtain the complete calibration curve. From here on, all antenna measurements are in terms of Channel 2 as reference. Separate calibration curves are needed for 3.8 and 7.1 MHz. It is important that straight lines connect plotted points. Any attempt to draw a smooth curve will lead to frustration and inaccuracies. Note that curves D and E do not give much discrimination between lead and lag in the region of ±30 degrees. However, the readings of the complementary angles will reveal the answer. Sometimes when measuring very small angles (±5 degrees) the complementary angle couldn't be obtained. This problem was resolved by installing coax of known electrical length in series with one input, then repeating the measurement.

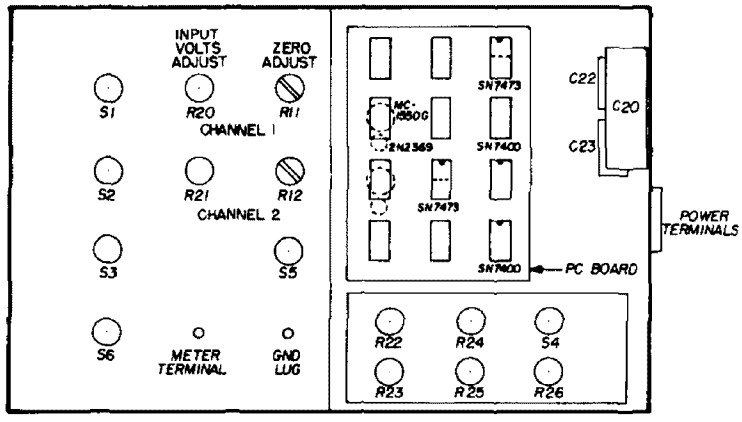


fig. 7. Component layout. Note that the MC1550G and 2N2369 devices are mounted in sockets, which in turn are soldered to the PC-board -mounted dual in-line sockets. (Other ICs are for experiments presently in progress.)

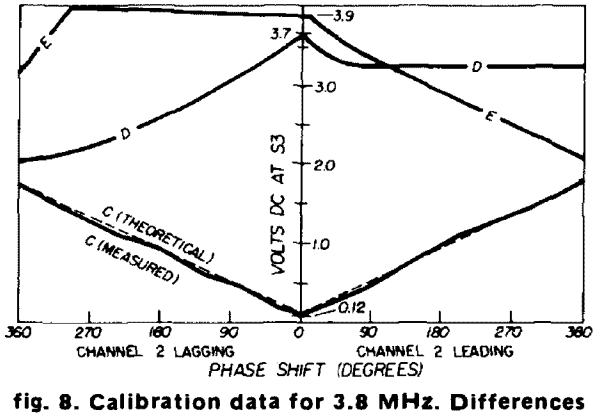


fig. 8. Calibration data for 3.8 MHz. Differences between theoretical and measured values of C are exaggerated to illustrate the need to connect points with straight lines. At 7.1 MHz the lines for C are similar, while those for D and E have the same shape but different values.

experimental notes

By feeding power to one antenna at a time, with the other antenna detuned, the pickup coils were positioned so that the voltage measured on the phase meter was identical for a given current flowing in the antenna. The results of typical calibration runs are shown in **fig. 9**.

To prove out the position of the pickup coils, the antenna elements were fed in parallel pairs, and nearly equal currents and small phase angles were noted. The coax cables to the antenna were of equal length but were only 0.9 wavelength electrically at 7.1 MHz. By increasing the coax lengths to 1.0 wavelength, unwanted

phase shift due to high swr was eliminated. With the three antenna elements active and the phase shifters and power dividers correctly adjusted, the f/b ratio (on the air) was 25 dB, which confirmed our design objective. Theoretical antenna patterns are shown in **fig. 10.**

acknowledgements and conclusions

This project was a team effort that included contributors over a wide area. John Ravenscroft, VE2NV, Montreal, provided many ideas and encouragement. Mr. Sam Tilden, also of Montreal, provided the data for the pickup coil design. Norman Tweit and Wes Vincent, Scottsdale, Arizona, contributed the basic Schmitt trigger circuit. George Oshiro provided the basic gating idea. Our colleagues at work provided many suggestions and our employer, the Aluminum Company of Canada, made test equipment available.

Preliminary tests have indicated that this instrument will work at frequencies to 10 MHz, and the limitation is probably due to the Schmitt trigger. The circuit was examined with the aid of a Tektronix scope, and it looks as if the flip-flops are another limiting factor. Operation beyond 10 MHz would probably mean using a different family of ICs, such as the EECL. A PC board would be essential.

If one were to substitute the higher-speed TTL devices, the result would probably be a more linear calibration curve, which would permit the output meter to be calibrated in terms of phase angle.

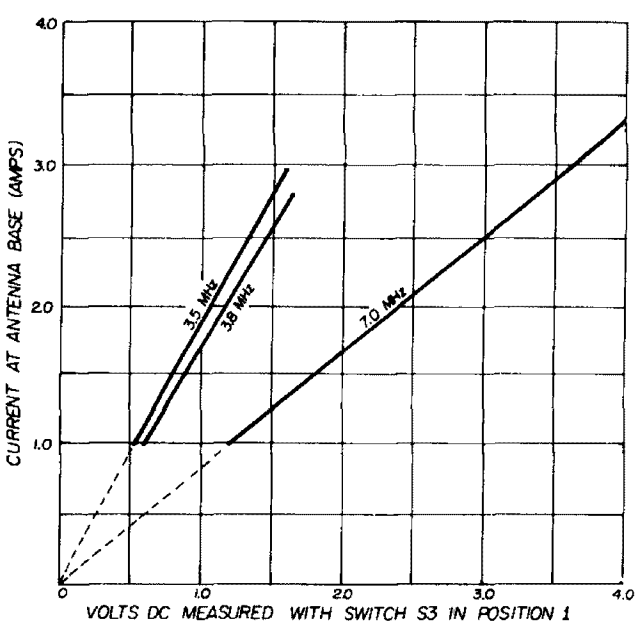


fig. 9. Pickup-coil calibration curves. The rf ammeter available was rated 0-4 A, and the meter scale was not calibrated in the 0-1 A range. Therefore the high-current points shown were used to prove system linearity and to correctly position the pickup coils.

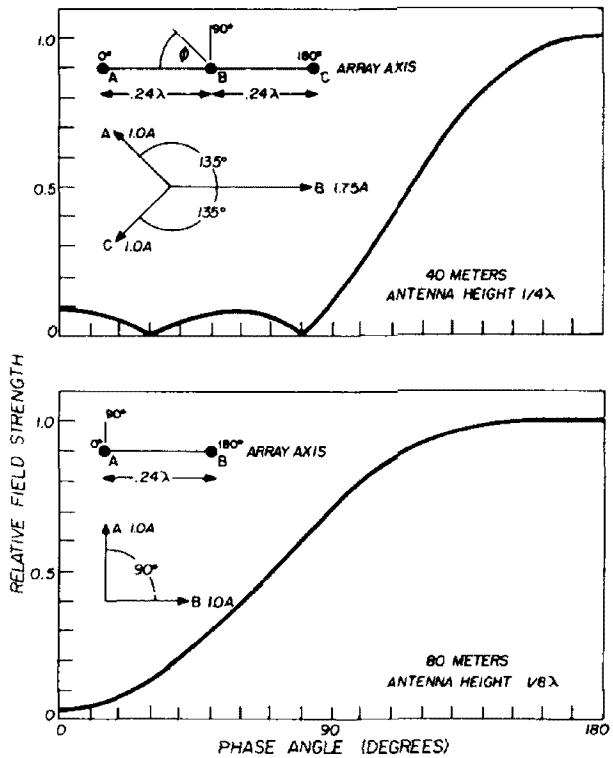
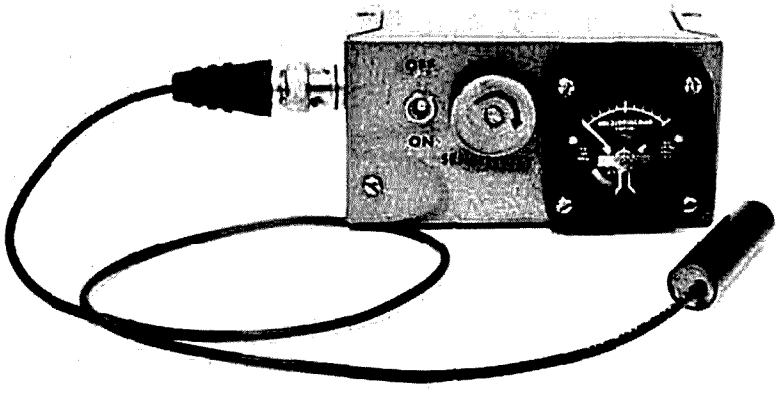


fig. 10. Antenna patterns. Data are shown in Cartesian rather than polar coordinates because null points are more easily identified. Data were plotted with the aid of a computer program (reference 2).

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5. Doyle Strandlund, W8CGD, "Amateur Measurement of R-jX," *QST*, June, 1965, p. 24.

ham radio



sensitive rf indicator

This high-gain amplifier circuit, when used with a DyComm Sniff-It, provides a sensitive rf indicator — the Snuff Box

D.L. Sprague, WB9DNI, 78 Paddock Street, Aurora, Illinois 60538

For some time now DyComm has offered an rf probe called the Sniff-It. This small black cylinder with a coax tail has proven to be quite useful in providing a dc current relative to an rf field. The manufacturer's instructions provided with the unit describe specific results of various applications, but do not provide general specifications such as frequency and power response. DyComm also has a very tight security rein on the contents of this midget marvel. So we approach the device as the classical black box. . . cylinder.

Art Householder, K9TRG, of Spectronics asked me to design an inexpensive relative rf indicator to be used with the Sniff-It, and the Snuff Box described here is the result of that request.

The Snuff-Box/Sniff-It combination is a portable rf level indicator useful in tracing problems in transmitters or receiver oscillator chains. By making the coax guide, also described in this article, leakage points in coax can be detected.

circuit description

The Snuff Box consists of a continuously variable gain dc amplifier feeding a five-volt meter. The gain can be varied from 1 to slightly more than 100. A μA741 frequency compensated operational amplifier comprises the active element. This device is both reliable and inexpensive. Because the 741 is internally compensated, an external frequency com-

pensation network is not required. The dual inline package (DIP) was selected for the prototype. If desired, the metal can package can be substituted. However, the pin configuration given in fig. 1 is for the DIP only.

The 741 IC is operated in the non-inverting mode to provide a relatively high input impedance. The input is protected by CR1, is a 5.6 V zener diode. R1 and R2 form a load for the Sniff-It with R1 acting to limit the input current when CR1 conducts. R2 was chosen by experiment as the approximate resistance for the desired power transfer. R3 and R4 form the negative feedback network which determines the gain of the op amp. Gain is represented by A

$$A = \frac{R3 + R4}{R3}$$

R5 and the 1-mA meter constitute a zero to 5 volt meter. The value of R5 is selected so that it's resistance, along with the meter's resistance, will equal 5000 ohms. There are numerous articles available which explain how to determine a meter's internal resistance. The 741 requires a bipolar power supply. I found the simplest to be two matched batteries.

construction

In this project, the meter is the prime consideration. Small 1-mA meters are very common junk box items, as this style was a favorite with military designers for many years. I recommend that the meter selected be of the ruggedized style. Many portable devices with non-ruggedized meters get quickly returned to the junk box!

Select the box for the meter. The prototype was built in a 2 x 4-inch Minibox. Layout is not critical. It's a good idea to use a socket for the μA741, as unsoldering 14 pins is a pain.

As purchased, the Sniff-It is not equipped with a connector. I suggest either a phone plug or a BNC type. Both are quick disconnect, and provide adequate shielding for low levels of dc current. However, the BNC is superior for high rf fields.

coax guide

Take two 3/8-inch cable clamps and cut as shown in the photograph. Pop rivet or bolt the two clamps together back to back, keeping them at right angles to one another. The Sniff-It will fit into one, leaving the second open to guide along RG-58/U coaxial cable.

operation

Let's examine the sensitivity of the Snuff Box. With R4 fully shorted, the

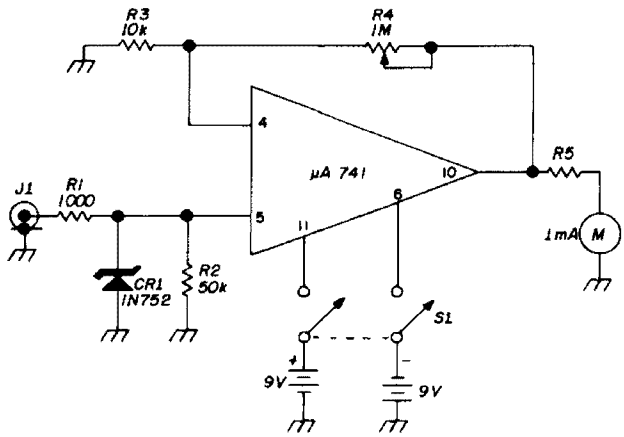


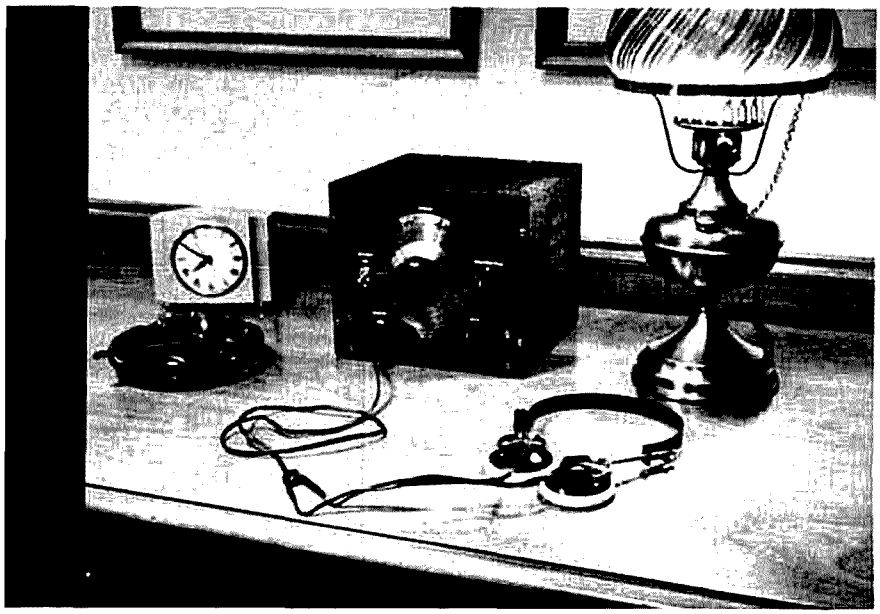
fig. 1. Circuit for the Snuff Box. DyComm Sniff-It is plugged into J1.

μA741 produces unity gain. Therefore, 5 volts across R2 will cause a full-scale deflection of the meter. This means that a current of 100 μA is being produced by the Sniff-It. Now, with a gain of 100, it will take only 1 μA for the highest setting of R4 to produce a full-scale reading. This sensitivity was sufficient for me to find the oscillating receiver crystal in my Regency HR-2.

DyComm's instruction sheet has many suggestions for the applications of the Sniff-It. The Snuff Box is ideal for most of these applications.

This unit is an invaluable aid in checking for rf on power supply leads, leaks in transmitter and antenna tuner cabinets, or for neutralizing detection. When peaking tuned circuits in a unit, the Sniff-It/Snuff Box is much handier than a grid dipper if you are just looking for maximum indications, since no tuning is necessary.

ham radio



simple regenerative WWV receiver

This low-voltage WWV receiver tunes from 4.7 to 15.5 MHz, and is powered by one 1.5-volt flashlight battery

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The receiver described here has a frequency range of approximately 4.7 to 15.5 MHz; this covers three of the WWV frequencies as well as the 20- and 40-meter amateur bands and several foreign broadcast bands. Its simplicity and economy are enhanced by the fact that it operates from a single flashlight battery.

I chose a regenerative detector because it will perform well with a-m, CW, or ssb. Regenerative detectors use positive rf feedback to obtain large amounts of amplification and very high Q in the input tuned circuit. As feedback is increased, Q and gain increase until a point is reached where the detector goes into oscillation. When oscillating, the detector provides its own bfo signal. All of the selectivity is obtained in the single front end tuned circuit, thus cross modulation and/or intermodulation interference is very low because unwanted signals don't reach the first transistor.

battery power supply

By designing the receiver to operate from a single flashlight cell, a consider-

able saving in battery expense has been achieved; also, battery changes are required much less frequently than if the popular 9-volt transistor radio battery is used.

Table 1 shows data extracted from battery performance curves in the Burgess Engineering Manual, 1967 edition. Only ordinary carbon-zinc batteries are shown. End-of-life voltage in each case was arbitrarily chosen to be 1.2 volts-per-cell, and all data is for continuous discharge at 70° F.

In comparing batteries, the important parameter is capacity in milliampere-hours; the 9-volt battery will supply 1 mA for 400 hours, but the D cell will supply 1 mA for 3200 hours, eight times as long as the 9-volt battery.

table 1. Carbon-zinc battery performance.

battery type	load current (mA)	life (hours)	capacity (mA-hours)
9 volt	1	400	400
AA	5	180	900
C	5	330	1650
D	20	160	3200

When you consider that the 9-volt battery may cost twice as much as the D cell, the cost-of-operation factor is about 16 to 1 in favor of the D cell. The above rational assumes the 9-volt circuit would draw the same current as the 1.5-volt circuit. If, however, the two circuits were designed to have the same output impedances, the 9-volt circuit could draw about six times as much current, making the cost-of-operation factor 96 to 1 in favor of the D cell.

headphone power

I used a subjective listening test to determine whether a 1.5-volt circuit could drive a pair of 2000-ohm headphones. The phones were connected to a 1-kHz audio oscillator, and an ac voltmeter used to measure the voltage across the phones. Table 2 shows the results. Since 300 mV rms is about 0.85-volt peak-to-peak, and since an amplifier having a 1.5-volt supply should be able to have an output swing of 0.85-volt peak-to-peak, there should be more than enough power available to drive the phones.

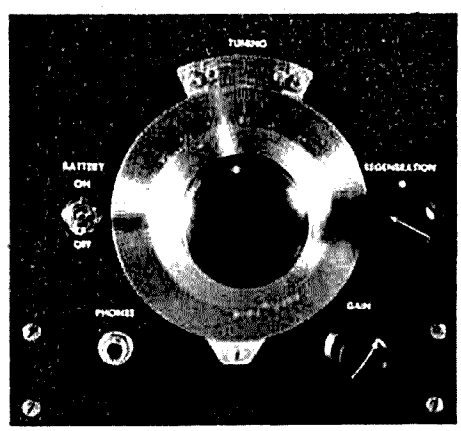
table 2. Headphone loudness vs supply voltage.

headphone voltage (mV rms)	power to headphones (microwatts)	observed loudness effect
2	0.002	detectable
30	0.45	comfortable
100	5.0	desirable
300	45.0	excessive

threshold howl

A common trouble in designing regenerative detectors is threshold howl. This is a loud audio-frequency oscillation heard when the regeneration control is advanced to the point of rf oscillation. It is caused by the circuit squegging or going in and out of rf oscillation at an audio rate. This is similar to the action of a superregenerative vhf detector which starts and stops its rf oscillations at a rate somewhere around 100 kHz.

According to Terman,[1] the tendency to have intermittent oscillations is reduced by decreasing the R-C time constants associated with the bias network. I had no trouble with threshold howl after I reduced the collector and base resistors of Q1 to the values shown in fig. 1. The



WWV receiver is built into an old Bud enclosure that was picked up at a swap-fest.

22-pF base coupling capacitor of Q1 should not be increased.

circuit

Fig. 1 is a schematic of the complete receiver. Battery drain is less than 1.5 mA.

Measure the dc voltage between the collector of Q4 and ground with a 20,000 ohm-per-volt vom or a vtvm. The audio gain should be set at minimum for this test, and 10 seconds allowed for the amplifier to stabilize after it is turned on.

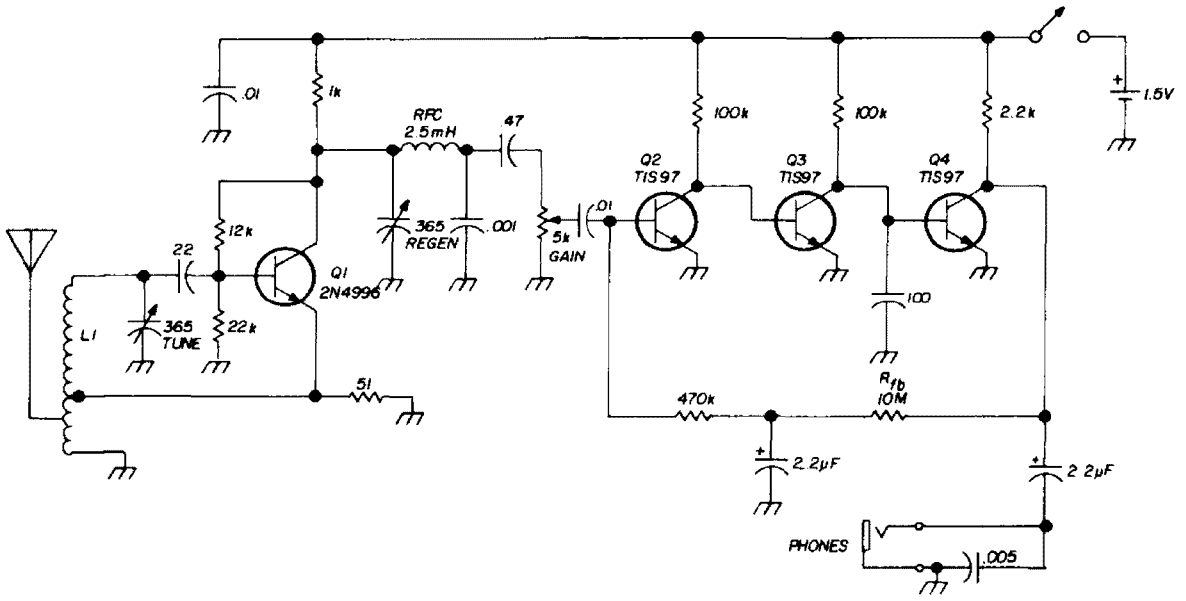


fig. 1. circuit diagram of the low-voltage WWV receiver which tunes from approximately 4.7 to 15.5 MHz. L1 is 3.8 μH (12 turns, 1½" diameter, 6 turns-per-inch coil stock).

If the voltage is not in the range of 0.75 to 0.85 volt, substitute different size resistors for R_{fb} until the voltage is about 0.8; larger value resistors will raise the voltage, and smaller ones will lower it.

The large coil stock was used to make it easy to experiment with tap positions. Antenna coupling is varied by means of a minigator clip from the antenna terminal which can be clipped to the coil at any point. The emitter tap is one turn from ground. A 51-ohm resistor from emitter to ground was necessary to ensure that oscillation could be stopped at all frequencies; it also stabilizes the circuit against antenna impedance variations.

There is nothing critical in the construction of the receiver, but, as usual, electrical stability is improved by mechanical rigidity, and short leads should be used in the detector circuit. The Bud C-973 cabinet, with chassis, and the National type N-2 dial were salvaged from an old vacuum-tube vhf converter I bought at a swapfest for a dollar. The photographs show how the tuning capacitor is mounted to a 1.5-inch aluminum angle and coupled through a flexible coupling to the dial. Not shown is the rf choke and audio amplifier which are mounted below chassis. All detector components are mounted to solder terminal strips, and the audio amplifier is assembled on a piece of Vectorbord.

Important considerations are the need for a speed-reducing dial because of the high selectivity and wide tuning range, and the desirability of a smooth action ballbearing capacitor for the regeneration control. The regeneration setting is rather critical, and anything you can do to make it easier is quite worthwhile. I used Calectro A1-227 variable capacitors for both tuning and regeneration.

operation

Use a fresh flashlight battery; old ones may have developed enough internal resistance to provide coupling between

circuits which results in undesired oscillations.

There will be about a 10-second delay after the battery switch is turned on before anything will be heard in the phones; this is because of the long time constant of the 10-megohm resistor and 2.2-μF capacitor in the audio amplifier feedback network.

The antenna tap should be set for the best compromise between sensitivity and selectivity. One or two turns from ground is about right for long antennas, but three or four turns can be used with short antennas. If selectivity is poor, move the tap toward ground.

Use only headphones having 2000 ohms, or more, impedance. Advance the regeneration control toward oscillation by increasing its capacitance (this turned out to be counter-clockwise with the variable capacitor I used). When the point of oscillation is reached, a sudden increase in the background hissing noise will be heard; if a signal is within a few kHz of the detector's frequency, a beat note will be heard.

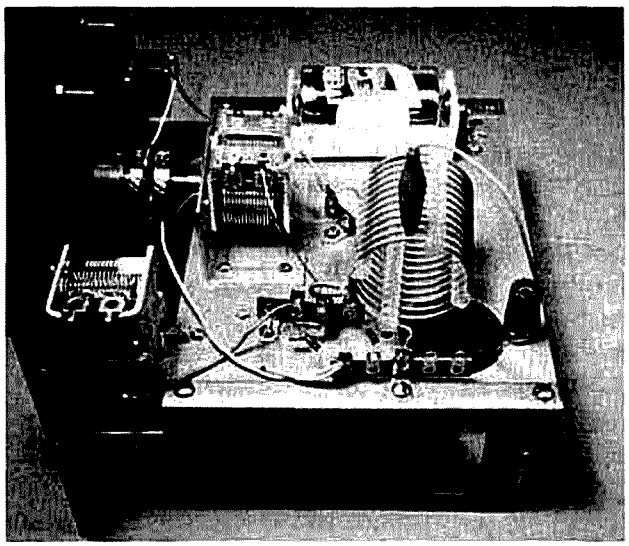
To tune in an a-m station, adjust the tuning dial and regeneration control simultaneously, keeping the regeneration just below the point of oscillation, until the station is clearly heard. There is interaction between the two controls, and some practice will be required to become proficient at tuning. Once the station is tuned in, it may be necessary to back off on the regeneration to eliminate distortion; the selectivity is so narrow at maximum circuit Q that sidebands can be clipped. CW signals should be tuned with the regeneration control set just above the point of oscillation.

With a tuning range of more than 10 MHz on 180 degrees of dial, it is a test of skill to tune in a ssb signal, but it can be done. The regeneration control should be set somewhat higher than for CW, and both tuning and regeneration carefully adjusted until the signal is clear.

High-powered amateur stations nearby can cause overload problems and be heard all across the dial, but this is not unusual, even with receivers of more sophisticated design. Moving the antenna tap toward ground will help.

conclusion

I hope the information above will help those who wish to build a simple general coverage receiver which is capable of good a-m performance. My reason for building it was to avail myself of WWV's accurate frequencies, time signals and propagation forecasts. I am well satisfied. Some will consider the critical two-handed tuning procedure a serious disadvantage, but I don't; things that require skill are more fun to do.



Construction of the WWV receiver. Audio amplifier is built on perforated board mounted underneath the chassis.

To my knowledge, this receiver is unique in that it can give good shortwave performance using a single flashlight battery power supply. Field Day enthusiasts and campers might consider adding a bandspread control for portable hamming. Regenerative receivers aren't as good as direct-conversion for CW and ssb, but they work.

reference

1. F.E. Terman, *Electronic and Radio Engineering,* 4th edition, McGraw-Hill, New York, pp 491-492.

ham radio

first wireless in Alaska

The frigid weather and frozen tundra were overcome to place a commercial wireless system in operation in 1903

Ed Marriner, W6BLZ, 528 Colima Street, LaJolla, California 92037

At the turn of the century the Alaskan gold rush was in full swing. Army forts had been established to look after the gold seekers, and a reliable communications system was needed. A wireless communications system was established, later to be called the Western Alaska Military Communication and Telegraph System, which eventually employed many radio amateurs who as civilians became speed operators on one of the country's fastest CW nets. They were called WAMCATS.

How did it all start? Historically, it was the first communications system on all of the West Coast of America. During the various Alaskan gold rushes, starting in 1898, towns and camps mushroomed overnight, with plenty of tin horn gamblers and other characters (as described in the poems of Robert W. Service) wandering from place to place. By 1900, better communications than messages by boat or dog teams became a necessity.

In 1900 Congress gave the U.S. Army Signal Corps money to improve wireless communication and charged them with the responsibility for a cable line to Alaska and the interior. This cable line was constructed under great difficulty for the poles had to be set into permanently frozen ground. This was the first of many factors which launched the Army into wireless communications.

In 1903 the U. S. Army had strung

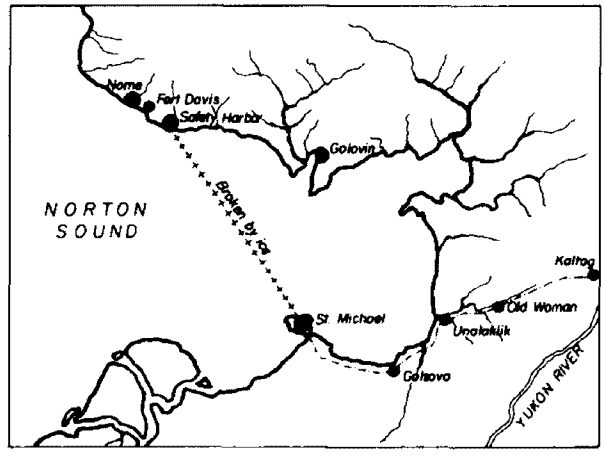
wires to St. Michaels, Alaska, and was faced with the problem of making it to Nome. This meant the choice of going around Norton Sound with a pole line under the worst of primitive conditions or laying a cable which would be carried away by ice each year.

The Signal Corps, by now, had used wireless telegraphy and decided that the last link between the stations at Nome and St. Michaels was to be a wireless one. The entire system consisted of 107 miles via wireless, followed up by 3883 miles of cable to Seattle. This same year, 1903, Dr. Lee DeForest experimented and exchanged satisfactory communication between Fort Safety, Alaska, and these stations which was the final report of the project which started back in 1899.

Leading up to the Alaskan Wireless Network was a report by the Chief Signal Officer who, in 1899, announced that the Signal Corps had devised a system of wireless telegraphy. This was the first publicly operated network in America. Improved in detail, it worked successfully over limited ranges between the harbor fortifications for which it was planned. Recognizing, however, that rapid advances were sure to be made by civilian experts, the Chief Signal Officer decided to adhere in this matter to his general policy. This meant that experimental work would be carried out by the Signal Corps only under those conditions where there wasn't any easy recourse to the commercial and industrial establishments of the United States.

Owing to repeated failure of several wireless telegraph companies to furnish a reliable and satisfactory system of wireless telegraphy in Alaska over a distance of 100 miles, the Chief Signal Officer decided in 1903 to have all existing systems examined with reference to their practical qualities. He decided to obtain by elimination, substitution or invention some system for Army use which would result in the reliable and successful transmission of messages.

The farsightedness of the Chief Signal Officer had already accumulated information and instruments which facilitated the solution of the problem. In addition to the systematic collection of all published data on wireless telegraphy, no efforts had been spared to supplement these by obtaining information from various inventors and experimenters. In addition,



Map showing the wireless circuit between Nome and Fort St. Michael, Alaska, a distance of 100 miles. This circuit was put into operation by the U.S. Signal Corps in 1903 to replace the cable which was washed out by ice every year.

the Signal Corps purchased sample instruments and installations pertaining to every system that seemed worthy of test where a title to the instrument could be obtained for a reasonable price. In this way they acquired essential parts of important systems. Experimental work in wireless telegraphy had also been done by Maj. Samuel Reber, George O. Squier and Edgar Russel, all Signal Corps people, but none of these officers were available for assignment to the work.

For experimental work in connection with perfecting a permanent plant, Capt. Leonard D. Wildman, a graduate of Stevens Institute of Technology, who had displayed resourcefulness in various phases of field duty, was selected. With full authority to call on Major Reber and Capt. Russel for advice, the accumulated data and instruments were turned over to

Capt. Wildman, who began his work by testing the capacity and efficiency of the Braun-Halske wireless transmitter, a duplicate of the plant which operated successfully during the German maneuvers. While this transmitter was not unsuccessful, its maximum capacity for transmission of messages over 63 miles was not entirely satisfactory.

In addition to determining the best type of field wireless apparatus suited to the Army, he was particularly charged with the setting up of a permanent wireless plant which should be able to work successfully over distances exceeding 100 miles.

The chosen field of operation was between Fort Wright, Fisher Island, and Fort Schuyler, New York, 97 miles apart, of which 20 miles was across land. The use of these forts was a factor in the national defense as it established a wireless system over which, in times of great disturbance, a message could be quickly exchanged between the outlying defenses of New York.

With these ends in view, experiments were carried on under Capt. Wildman's personal supervision during the latter part of the year with instruments purchased from the Lodge-Muirhead Wireless Telegraph Company of Great Britain, the Brau-Siemans-Halske Wireless Telegraph Comapny of Germany, the National Electric Signal Company of Washington, D.C., and from the DeForest Wireless Telegraph Company of New York.

In addition to the instruments furnished by the above companies, experimental apparatus was purchased by the Chief Signal Officer from time to time. Comparative tests were also made between all receivers, responders and coherers on the market, as well as with many different types of special equipment.

Not withstanding the popular idea that wireless telegraphy over great distances was an accomplished commercial fact, *none* of the systems investigated proved satisfactory for Army use. (Have times changed?) Wireless telegraph systems seemed to have been developed by their inventors in the laboratory for their own use rather than to electrical and mechanical standards in which a reliable piece of equipment could be placed on the commercial market.

After an investigation of the existing systems, Capt. Wildman formulated, with the approval of the Chief Signal Officer, the following changes as being necessary for practical military uses:

1. Eliminate the necessity for an absolute electrical ground.

2. Construct all parts of the apparatus so that in case of the failure of any part, that part can be replaced without elaborate machinery by intelligent unskilled labor. Even in those days they had trouble getting technicians.

3. Replace all adjustments which require a knowledge of mathematics, or experience in manipulation by lettered dials or definite switch positions so that highly skilled operators would be unnecessary.

4. Reduce the necessary height of the antenna wires.

5. Produce a receiver which would not only receive the message intended for it, but which could, by adjustment, also receive any electromagnetic wave.

6. Eliminate disturbances due to atmospheric or static electricity.

7. Avoid, as far as practicable, all dangerous high-potential currents at points where there was a possibility of danger to employees.

8. Provide devices which would protect the instruments and machinery from destructive potentials.

9. Avoid, as far as possible, the use of patented devices and the consequent payment of large royalties.

10. Devise a system which could be easily transported in time of war and which would be capable of transmitting messages under all climatic and topographical conditions.

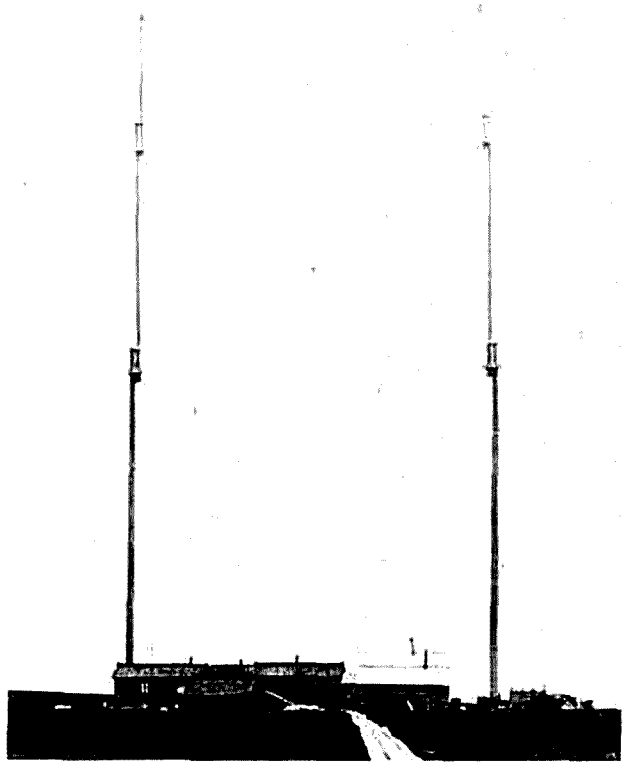
Extended tests were made on the DeForest System which in its original and earlier forms was successfully operated during the Army and Navy maneuvers in 1902 on Long Island Sound. This was the first time the Signal Corps applied wireless telegraphy to military purposes. During these tests the DeForest system barely covered the Schuyler-Wright course. Under the most favorable conditions a signal could be exchanged successfully, but when there were nearby interfering stations, signal exchange was impossible.

Undismayed by the situation Capt. Wildman applied himself to the problem of supplementing and improving the system, to the point that he solved the problem as far as the needs of the Signal Corps were concerned. His improvements were formulated and patented equally in the interests of the United States, to whom the patents were assigned, and of the inventor.

Although the Signal Corps system was not perfect, it was better than any system previously tried. It was not absolutely unbreakable, it could not be operated by men of a low order of intelligence, and it was not entirely free from interference from nearby stations, nor could it be operated during heavy thunderstorms. Although, from the looks of the tests, it did not meet many of Capt. Wildman's specifications, the U.S. Army was eager to give it a try in Alaska. A good substitute for a good electrical ground had been found, the operations and adjustments were simple enough, there was only one place within reach of the operator where there was a destructive potential, and the equipment was decided to be repairable enough.

Messages during the tests were sent daily in great numbers over the ninety-seven mile path for five weeks without any apparent deterioration in apparatus or machinery. The experiments furnished a large amount of accurate and valuable data on placement of antenna poles, their rigging, construction, dynamos and their design, transformers and their durability in moist weather, induction coils and their action, the various methods of tuning the antennas to each other, and to the closed oscillating circuits by which they were fed. The tests proved very valuable.

With the exception of the DeForest receiver, the Signal Corps system had no patentable devices other than those invented and designed by the officers and enlisted men of the Signal Corps. As usual, the general public recognized that they had been taken in by the extravagant claims of wireless telegraph experimenters. The public would have to wait for experience alone to prove whether the devices adopted by the Signal Corps performed properly when transferred from a temperate climate to places like Alaska where they had to be operated and maintained by unskilled labor in an unfavorable environment.



Old wireless telegraph station at Fort St. Michael, Alaska. (Photo courtesy U.S. Signal Corps)

It appeared for certain, however, that for Army uses this system was better than anything previously available on the market. Time might disprove the utility of

some of the features which seemed promising, but the Army felt it was an advance over other wireless systems then in use.

Capt. Wildman was about to tackle the job of installing a communications system in the cold wilds of Alaska, and acknowledged his indebtedness to Major Reber and Capt. Russel for valuable advice and assistance. Special commendation is due Capt. Wildman for the persistent and skillful manner in which he contributed to the efficiency of the Army in perfecting the Signal Corps system of wireless telegraphy. The improvements were largely his own devices. Two patents were obtained by him for wireless inventions and were assigned to the government. He went on to Alaska, completed his task and wrote the following report to the Chief Signal Officer in 1903:

wireless telegraphy

"The system of Wireless Telegraphy devised by the Signal Corps of the Army in 1899 has been improved in details, but its range of operation is limited. It was deemed advisable to stop experimental work along these lines pending the development of this science by experts in civil life.

In 1901, however, it became a matter of practical importance to the Signal Corps to establish wireless telegraphy over extended distances. A contract was made looking into the establishment of the wireless telegraph by the Fessenden system across Norton Sound from Nome (Fort Davis) to St. Michael, about 110 miles. The contractors failed, however, to make the installation and the contract was revoked.

In view of the failure of the contractor to install the wireless system across Norton Sound, Alaska, and in order to meet the desire of the Commanding General, Department of Columbia, for telegraphic communication with Fort Davis, the Signal Corps took up this problem and is now engaged in an effort to install a system that shall work from St. Michael to Safety Harbor, near Nome, Alaska, a distance of about 105 miles. Experimental work with separate and composite systems is being carried out in Long Island Sound by Capt. L.D. Wildman, Signal Corps, with a view to eventually be working between Fort Schuyler and Fort H.G. Wright, a distance of 105 miles, and of which about 10 miles are lowland. For this purpose masts 140 feet high have been constructed, and completed in order to make final tests. Captain Wildman now awaits special motor dynamos and transformers.

Meanwhile to the delay, installations of masts and antennas are now being made at Safety Harbor and St. Michael, so that whatever system proves satisfactory in Long Island Sound, can be utilized in Alaska by 1904 with the suitable sending and receiving apparatus. At both St. Michael and Safety Harbor the permanent plants are now in process of transportation and erection. There are to be at each station two triple masts 200 feet high, between which are to be a suspended fan shaped antenna, consisting of 125 copper wires one-foot apart. The motor power is to consist of a 5 hp gasoline engine and a 3 kW motor dynamo, 60-cycle alternator. At one station will be a transformer, stepping up from 500 to 20,000 volts, and at the other, stepping up from 500 to 25,000 volts. The large Muirhead receivers, which now seem to be the best available type, are to be utilized in this work unless meantime other experiments produce something superior.

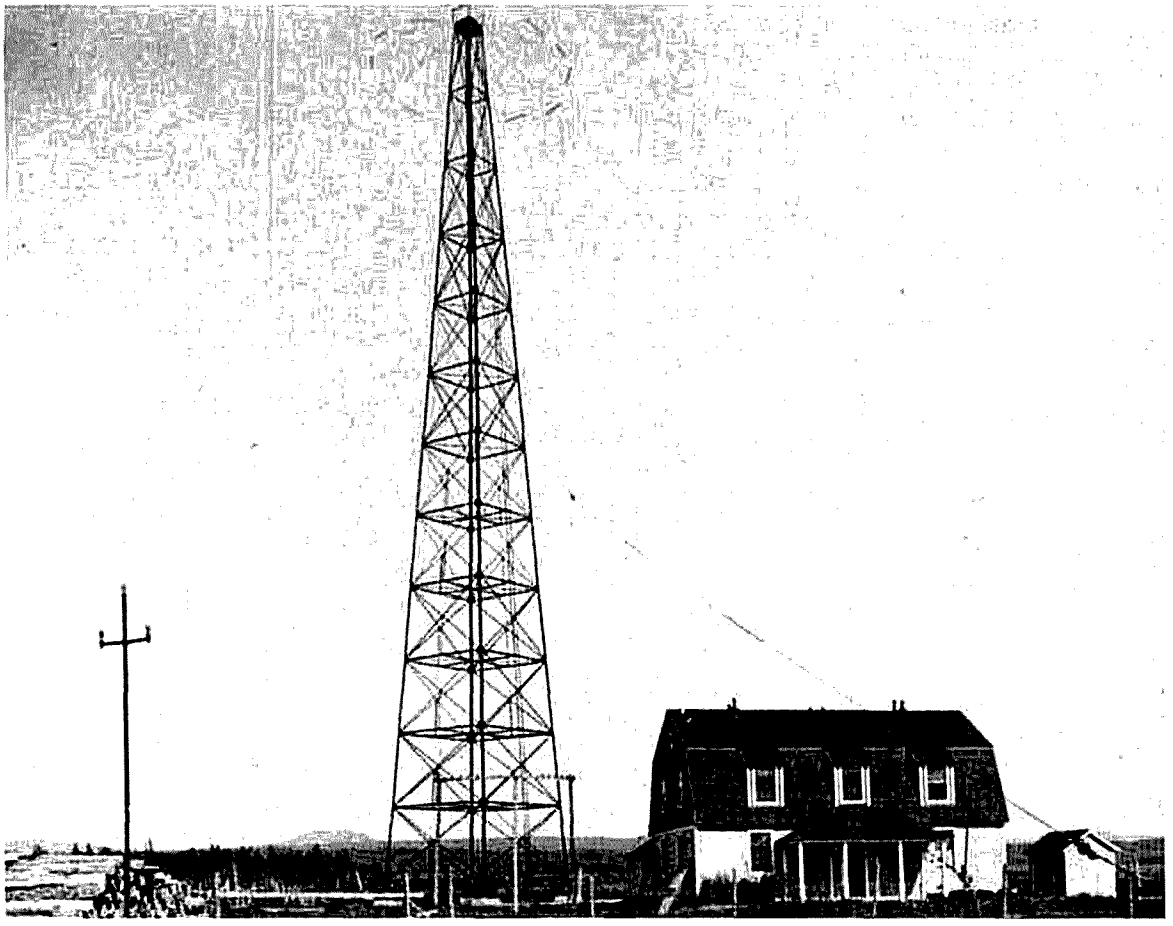
Another contract was made with the American-Marconi Wireless Telegraph Company to establish wireless communication between two points in the Tanana Valley where great difficulties were expected in constructing an ordinary telegraph line and in maintaining it satisfactorily, the contract locating the connecting of two points about 164 miles apart with an intermediate station should the Marconi company so decide. It was hoped that this installation would be made by October 1902, but the contractors were not able to install the system last year. They were at work

during the summer of 1903, but to this date no success has been reported. It has, therefore, been necessary for the Chief of Signal Officer of the Army to direct the efficient maintenance of the land lines in the lower valley of the Tanana, such action being imperatively necessary in view of the failure of the wireless installations in a reasonable time.

cycles. This runs through a two-kilowatt transformer, which steps up the voltage across the spark gap. Messages are received by the telephone and DeForest responder."

Report to the War Department, Washington, D.C., October 3, 1903

"While communication is now heard



United States Signal Corps station FD at Nome, Alaska, 1903. (Photo courtesy U.S. Signal Corps)

As we stated in the last annual report, the DeForest system of wireless telegraphy was utilized during the Army and Navy maneuvers on Long Island Sound. This year the same system has been used to replace a broken cable in New York Harbor, between Forts Wadsworth and Hancock, and it has worked most satisfactorily over this distance of 12 miles. In this system a motor dynamo of one kilowatt capacity, driven by the power of the post plant at 110 volts, produces an alternating current of 500 volts at 60

regularly by telegraph between the civilized world and the Yukon Valley westward to St. Michael, yet restoration of communication with Nome has so far proved impractical. The cable between Nome and St. Michael was so badly injured by ice, some 40 miles of it having been carried away, that its repair meet the urgent recommendations of the Commanding General, Department of Columbia. Efforts are being made, with prospects of success in 1904, to establish communication by wireless telegraphy

between St. Michael and Nome across Norton Sound, Alaska, a distance of 108 miles."

October 4, 1904, from Capt. Wildman

"The Signal Corps wireless station at Nome could communicate with a similar station on the Kamchatkca Coast, but the infertile and sparsely inhabited country thence to the nearest Russian station of Nikolaevisk would render any such enterprise unlikely should it be suggested. As has been stated in previous annual reports, efforts to establish a wireless system across Norton Sound and in the valley of the Tanana, awarded to different companies under public proposals, failed entirely.

The contract to establish communication was abandoned in its primary stages in one case, while in the other the efforts proved fruitless after two or more years trying to construct a permanent line through that section, and it became necessary for the Chief Signal Officer of the Army to undertake through the officers and men of his Corps, a wireless installation across Norton Sound, and work commenced along two lines: First, to install those available and then perfecting the system later. Second, as the short navigable season of four months in Norton Sound rendered it impracticable to carry on experimental work in Alaska, it was decided to establish two stations in connection with the coast defenses of the United States, where they would have a permanent value, and after devising a successful typical plant, transfer its sending and receiving apparatus to Alaska."



Inside the Signal Corps station FK at Circle City, Alaska, 1909. (Photo courtesy U.S. Signal Corps)

The arrangements for the temporary plant were made under the general directions of Major Russel, whose special and important duty in connection with the Alaskan cable installation left him but scant time for arrangements, whose executive duty must necessarily be carried out

2,000 miles away by an assistant not under his personal observation.

In the late summer of 1903 the Norton Sound base was established. At Safety Harbor and St. Michael there were built portable houses, in which were installed engines, batteries and wireless instruments, supplemented by two masts at each station 210 feet high, between which were suspended fan-shaped antennas. These poles, the highest ever erected on the Pacific Coast, and the antenna were installed through the resourcefulness and professional skill of Mr. R.D. Ross, a civil engineer employed for this purpose.

Unfortunately, part of the wireless material failed to reach St. Michael because the steamer it was on, the *Meteor*, was disabled enroute. First Lt. A. T. Clifton, with a selected force of signal-men familiar with wireless work, jury-rigged instruments through which meager wireless signals were exchanged during the winter.

Meanwhile, experimental work was carried on in Long Island Sound by Capt. Wildman with separate and composite systems. He eventually devised a composite plant, originally based on the DeForest system, but largely modified by inventions of his own. This plant worked with great success between Fort Wright and Fort Schuyler New York. The transfer of the wireless equipment from Long Island Sound to Norton Sound was accomplished by Capt. Wildman during the next summer. The method of installation was such that the installation was easily made.

Capt. Wildman, at St. Michael, and Sergeant Treffinger, at Safety Harbor, installed their respective systems in less than two days. Capt. Wildman reported that the wireless material was landed and delivered at St. Michael by noon of August 4, and said:

"At 9 o'clock AM on the 6th, complete messages were exchanged, and the telegram from me at Safety Harbor was released and set forward. No serious trouble of any kind was experienced and every part of the machinery worked in a perfectly satisfactorily manner. Since that time we have been pushing the machinery about 20% overloaded in order to see if it could be broken down. The signals are fine and louder than I have ever heard them at either the stations when at Schuyler or Wright. The operators have no difficulty in reading the messages while the relay is working in the same room and with the engine running in the next room and men walking about and talking in an ordinary voice anywhere in the house."

table 1. Callsigns of some of the early U.S. Army Stations in Alaska.

FB	Fairbanks
FD	Nome
FE	Mouth of the Yukon
FG	Fort Gibson
FK	Circle City
FM	Ft. St. Michael
FP	Petersburg
FQ	Ft. Egbert
FX	Ft. Worden

On August 17, 1903, the Nome station was thrown open for commercial business with the rest of the world, and the wireless section of the Alaskan Telegraph System was an everyday adjunct of the electrical appliances of the twentieth century. It daily transmitted the entire telegraphic business of the Seward Peninsula. In one afternoon 5000 words were exchanged between Safety Harbor and St. Michael.

Through the professional skill of Capt. Wildman and his subordinates, the Signal Corps had started operating the longest wireless section network of any commercial telegraphy system in the world. Some of this early equipment was still functioning as late as 1922, in Craig, Alaska. The old 3-kW transmitter was still run by the gas engine, and the transformer, glass-plate condensers, straight open spark gap with cup-like electrodes and plain helix were still going after all those years. Wireless proved itself easier to maintain than telegraph wires.

ham radio

how to make your own printed-circuit boards

Complete directions on how to make your own printed-circuit boards — a practical AFSK generator is included as an example

Bert Kelley, K4EEU, 2307 South Clark Avenue, Tampa, Florida 33609

Today's solid-state components are small with terminals close together. Often the mechanical support and the terminal pins are one and the same, and printed-circuit construction eliminates the difficulties of wrapping wires around IC pins only a tenth of an inch apart without shorting adjacent pins. PC boards also solve the mounting problem. Wiring then becomes very easy, high parts density is possible with little effort, and the printed-circuit assembly is rugged — not likely to fall apart under vibration or use.

The main requirement is to carefully plan the parts layout on the board. This may be a different approach from the old vacuum-tube days when wiring began when all the tube sockets were mounted, often without much planning. But planning actually makes the wiring go faster because once the board is etched, the resistors, capacitors, and ICs almost literally fall in place and the project is finished in a very short time. Making your own circuit boards is not at all difficult if you assemble the material and tools I suggest and follow the following instructions.

tools

Many hobbyists already have the few tools required (see fig. 1). First of all, you

need a tool to cut the circuit board. Phenolic board is brittle and splits when cut by a hacksaw unless it is carefully clamped. If you do much construction work, you will find that a saber saw does a better and faster cutting job and is useful for chassis metal work. You will also need a quarter-inch drill, preferably variable speed.

The drills shown will handle most of the board drilling you will do. I use a no. 60 drill for all IC and transistor leads. IC pins will fit into number-77 holes but I don't recommend them as you might want to remove the component later, and the tight fit only makes the job more difficult. The rest of the drills needed will be in a small drill set. This will cover holes for PC board mounting, transformer mounting, 2-watt resistors, PC pots and the like.

The awl is used to transfer hole markings from a circuit template to the circuit board. This is shown in one of the photos. Crocus cloth is used to smooth the copper side of the board after drilling before application of resist ink.

The soldering iron should have a small tip. A device is shown for removal of solder in case you get solder in the wrong places or want to remove a component from a board. A similar tool is sold by Radio Shack (no. 64-2085). You can also remove solder by using a short length of fresh coax braid as a blotter, but since there is some risk of overheating the circuit board it is better to equip yourself with the proper tools.



fig. 1. A few simple tools are required for building your own printed-circuit boards.

etching stand

For best results an etching stand is necessary (fig. 2). It agitates the solution and gives a uniform, faster etching job. Fig. 3 shows one large enough for almost any amateur project. It consists of a 1/8-inch aluminum plate (which may be cut from an old rack panel) upon which the plastic etchant tray is placed. The aluminum plate is supported by two hinges at its center of gravity and is tilted by action of a small electric motor, fig. 4. The base is cut from plywood. The motor may be from four to eight rpm, such as the motors used to operate small animated signs. They can be found in stores listed in the *Yellow Pages* under "Display Fixtures." Try to locate one with the mechanical linkage attached that fits the small diameter shaft. The detailed drawing and photos show construction and mounting of the motor mechanism.

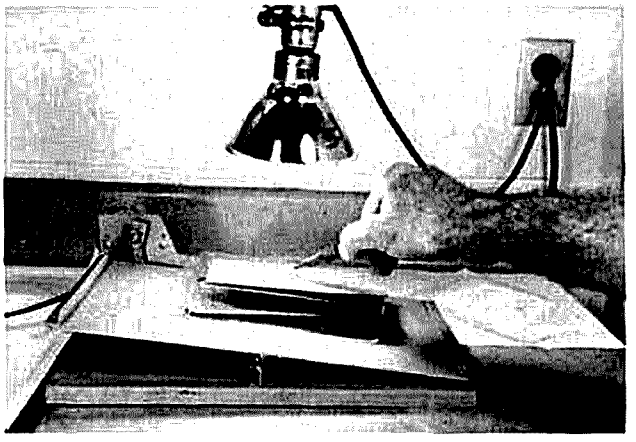


fig. 2. An etching stand is required for best results. Complete construction details are shown in fig. 3. Floodlamp is used to heat the etching solution.

Fasten small aluminum angles or strips to each end of the tilting plate so that large trays will not slide off the stand. Use plastic trays slightly larger than the board you intend to etch; this conserves etchant. Pyrex baking dishes and photo developing trays make good containers for the larger jobs.

A strong light suspended over the etching tray will speed the etching by keeping both the board and etching solution warm. I use an ordinary clamp-on fixture with a 150-watt floodlamp temporarily clamped to a cabinet above the sink. With this setup a board will be completely etched in less than ten minutes.

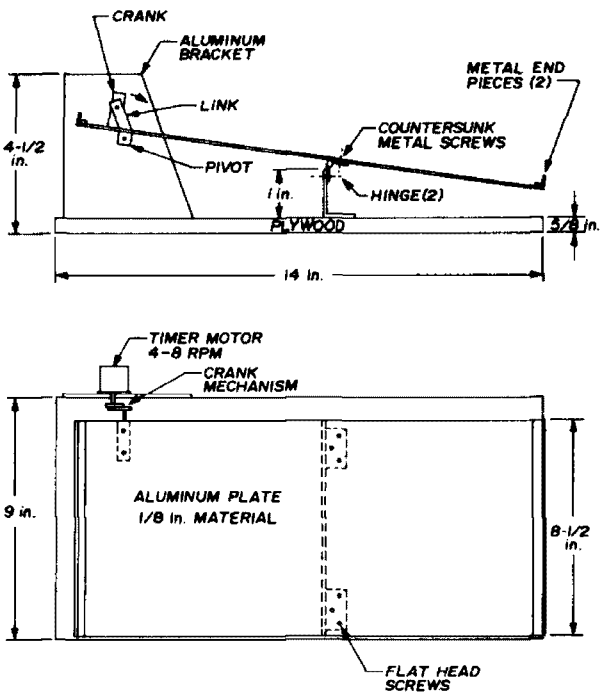
etching materials

The photograph shows the etching materials that I use (fig. 5). There is a variety of materials available, and this is a matter of preference. I find etching powder more convenient to store and keep several bags on hand. Once mixed, the powder should be used immediately as it will not keep it's etching strength. Use only as much as you need and seal the plastic bag so the etching powder will not absorb moisture from the atmosphere. The small bag shown in the photo will do about three 3 x 6-inch boards and costs $1.00. It is sold by Amidon Associates.*

Of the various substitutes for resist such as paint, marking pens, and masking tape, masking tape is the most reliable and covers large board areas well. Any substitute should be tested on a sample board. I use a Kepro RMP-700 resist pen which has a fine enough point to mark IC pins easily and effectively resists the action of the etchant.

The common brown phenolic circuit-board stock, copper clad on one side, is satisfactory for anything except preamps for vhf or uhf. Epoxy boards should be used where better insulation or greater strength is required. Unfortunately, epoxy boards are relatively hard on drills. Double-sided boards are not needed for most projects and only make layout and component removal difficult. However, double-sided boards are used where an excellent ground plane is necessary.

The circuit-board template is made on a 10-line-per-inch ruled pad. Most electronic components have lead spacings that are multiples of 0.10 inch. One-half watt resistors require 5 spaces, IC pins use



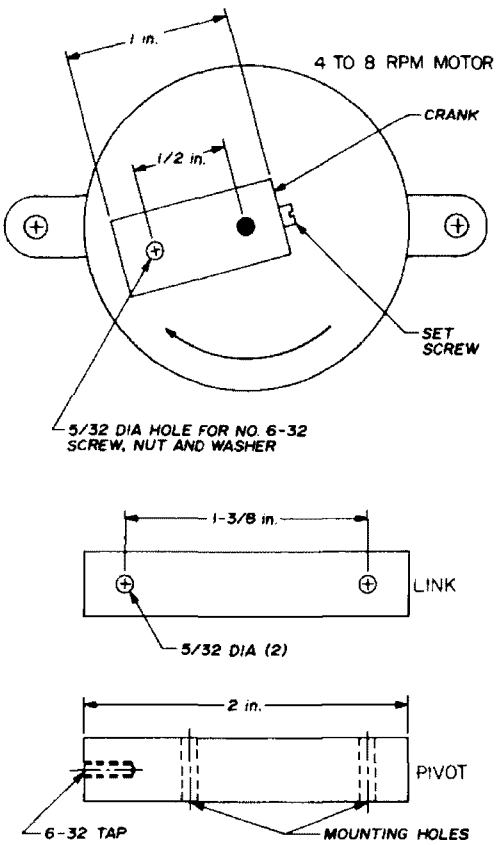


fig. 3. Construction of the printed-circuit etching stand. Sheet of 1/8-inch aluminum can be made from an old rack panel. Motor-drive system is shown on the right. A closeup photograph of the motor drive is shown in fig. 4.

*Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607.

1 space, etc., so layout is facilitated by using this paper. It may be obtained at any stationery store.

procedure

All components for the project should be on hand so they can be checked for size when layout begins. You can save here because the layout may be made to fit the parts you have, rather than in reverse. When you must buy new parts get small ones. There is no advantage, for example, in getting large 600-volt paper capacitors when small 50- to 100-volt units will be more than adequate and save much space. Always pre-test the parts where this is practical. Even experienced counter personnel at supply houses will sometimes misread the color code on resistors. Such errors can be difficult to detect after the parts are mounted on the circuit board.

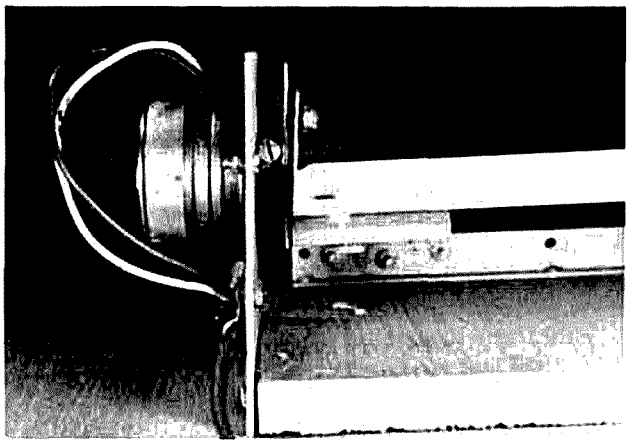


fig. 4. Small electric motor is used to agitate the etching stand.

Lay out the parts on the board roughly as they will appear in the schematic diagram (fig. 7.). Then, consider the layout, move components around in an effort to shorten leads, eliminate possible feedback, space parts more evenly, provide for external connections and reduce trail crossovers. This is the most important phase of the layout procedure. Where possible, refer to the author's original layout or photograph. If this is not given and the project is too complex to visualize as a complete unit, place all the components on top of the proposed circuit board and measure the resulting area. This will give the size of the board needed and the size of the enclosure.

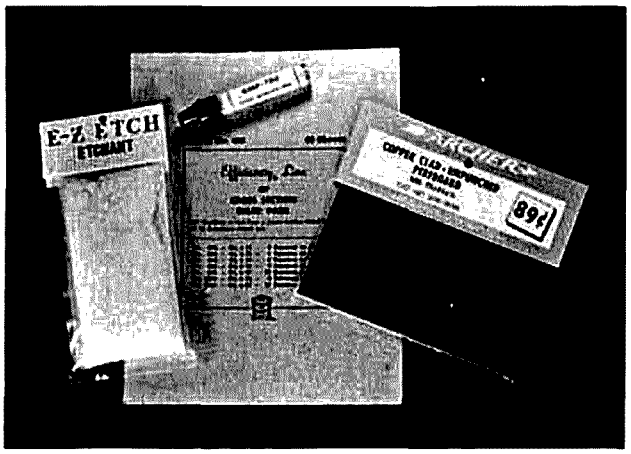


fig. 5. Etching materials are available from a variety of sources.

Now, select the enclosure, if you have not already done so. Miniboxes are available to fit the majority of small electronic projects. Mount the unetched board inside the enclosure. A mounting method that works well is to use threaded brass spacers like H.H. Smith no. 2372 at each board corner. Another simple mount may be constructed by running extra nuts on long 4-40 or 6-32 machine screws and using these for spacers.

layout

Now, make the detailed layout. Every hole in the circuit board should be shown in this layout and every circuit trail should be penciled in as it will appear on the finished product (fig. 9). It helps to

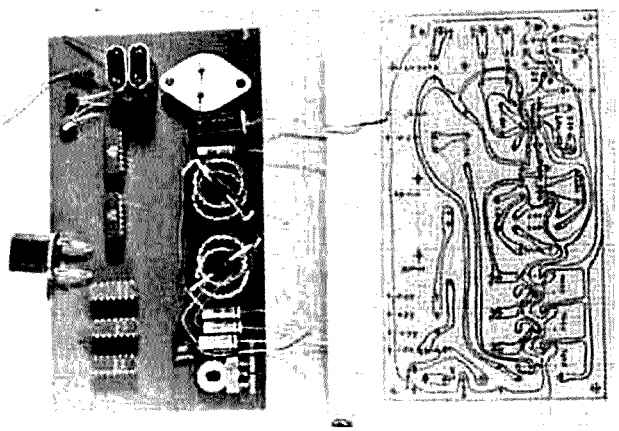


fig. 7. Before starting the printed-circuit layout, lay out the components on the board roughly as they appear in the schematic diagram.

overline a copy of the schematic so that nothing is overlooked. Try to keep trails as direct as practical, and have a minimum of crossovers. On your first few boards you may find it necessary to scrap the first effort and start over, if, during the late stages of the layout a better arrangements for different packages of the same type IC.

A sample layout is shown for an AFSK generator intended for radio teletype use (**fig. 6**). The tones are crystal controlled and provide low distortion output. It is not difficult to construct and illustrate

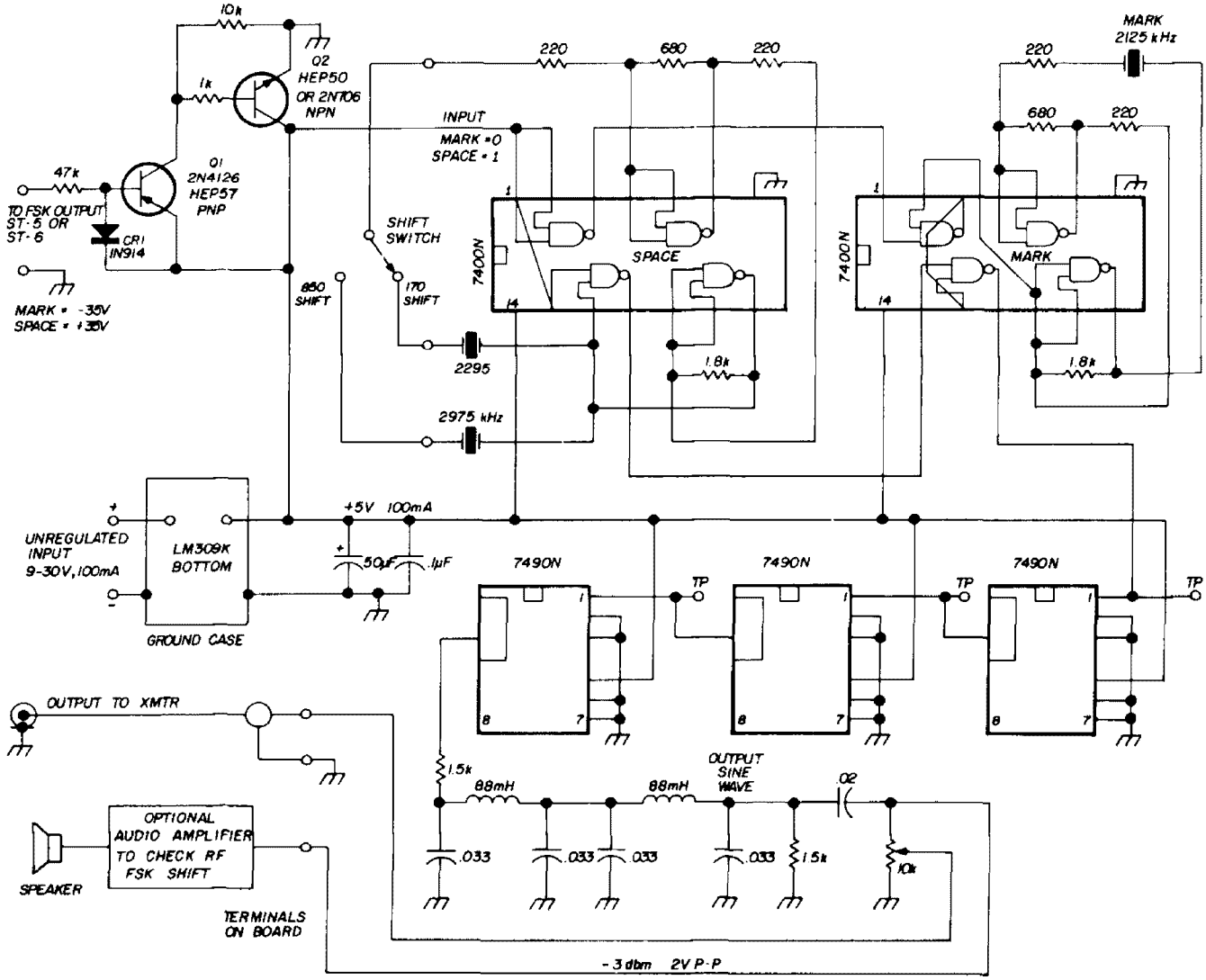


fig. 6. Circuit of RTTY AFSK generator used to demonstrate printed-circuit construction.

approach occurs to you. This is the time to make changes — when they can be easily done.

Keep in mind that the components will be mounted on the board with the leads facing you as you make the template. The pin arrangements for ICs are given as top views, while you will be laying out the pin wiring as viewed from the bottom. Also, be sure you have the pin layout for the particular IC you are using because there often are other pin the principles of home workshop printed-circuit construction. It will be seen that the board layout closely follows the schematic and only two jumpers have been used. A common ground has been drawn around the outside edge of the board and the metal corner spacers will continue this connection to the case.

A few kinks are useful. The crystals may be made to plug in by using contacts salvaged from crushed miniature tube sockets. They are then spot soldered to

the circuit board. Faucet washers make an effective mount for the 88-millihenry inductors. Thread a long 6-32 screw through one. The taper of the washer will center the screw and the resilient washer material will prevent coil damage.

An excellent way to connect to circuit boards is also shown. Force 1/4-inch lengths of no. 12 wire in 5/64-inch holes in the circuit board and etch trails to these points, under the board. These make secure test points or external connection points at almost no cost.

After checking, the completed template is taped to the copper side of the circuit board and the hole centers are transferred to the copper-clad board with an awl, using light hammer blows, **fig. 10.** When all holes are marked in this way the paper is removed and all marks drilled with a no. 60 drill. Larger holes are then located and re-drilled to size. Remove all drilling burrs with the crocus cloth so the board is smooth and ready for the resist pen. Avoid coarse sandpaper as it will scratch the board and make complete application of resist difficult. The etchant will later eat through these scratches causing hairline opens in the circuit trails.

The final step before etching is application of resist ink. This only takes a few minutes since you follow the trail layout you have previously marked on the template. Make the resist trails solid and dark.

The etching stand is set up and the lamp secured in place over it. Mix the required amount of etchant in hot water and float the board, copper side down, in

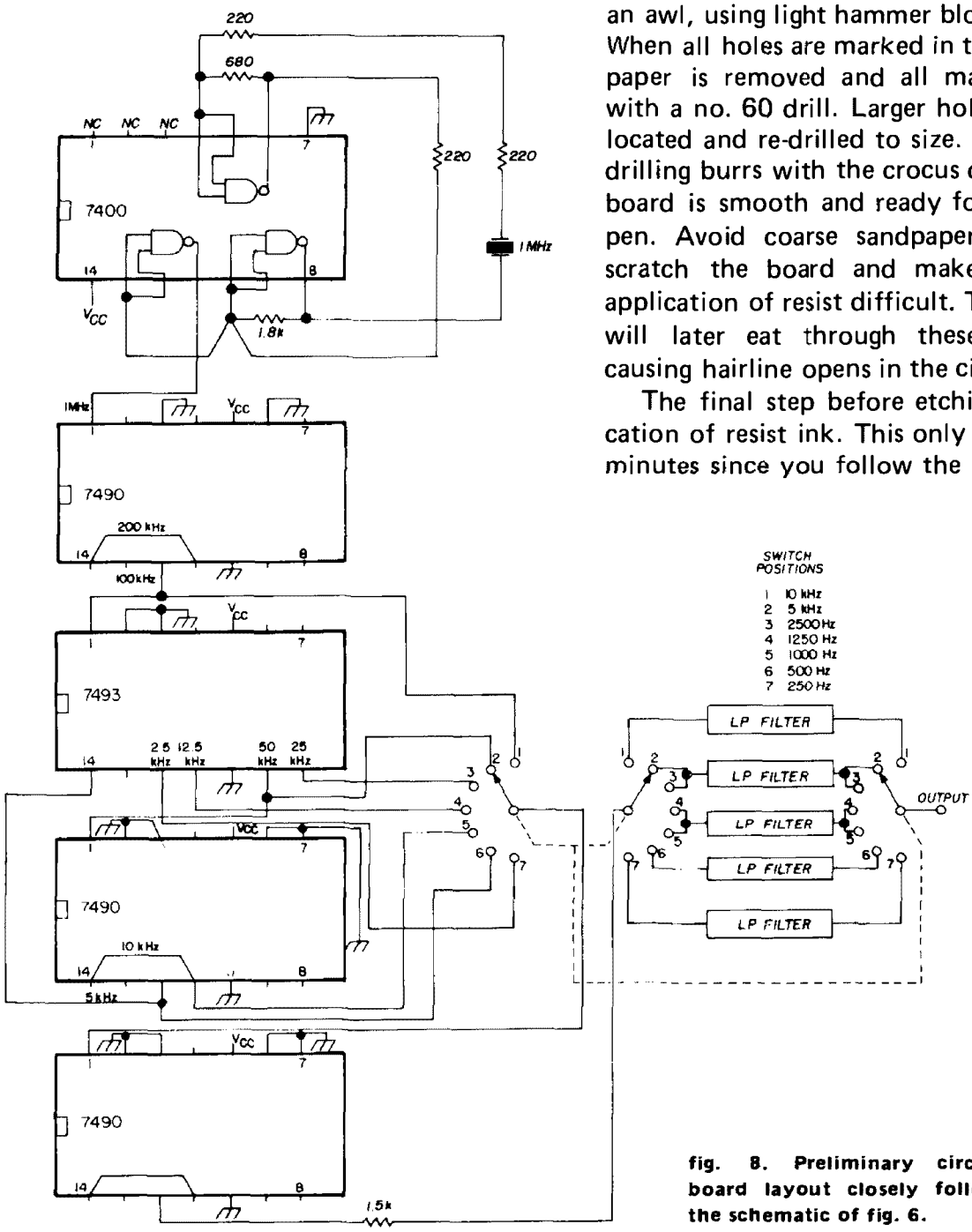


fig. 8. Preliminary circuit-board layout closely follows the schematic of fig. 6.

this solution. The board should be frequently lifted during the etching process to free accumulated air bubbles, fig. 11. One way to do this is to attach a small enameled wire through existing holes near the center so that it may be grasped by long-nose pliers. Do not overetch. As soon as the board is finished, remove it from the tray, turn off the light, and wash the board with cool water. Scrub the resist ink off with steel wool. The used etchant will be greenish in color and hot. Handle it carefully, dump it down the drain, and immediately flush with plenty of water.

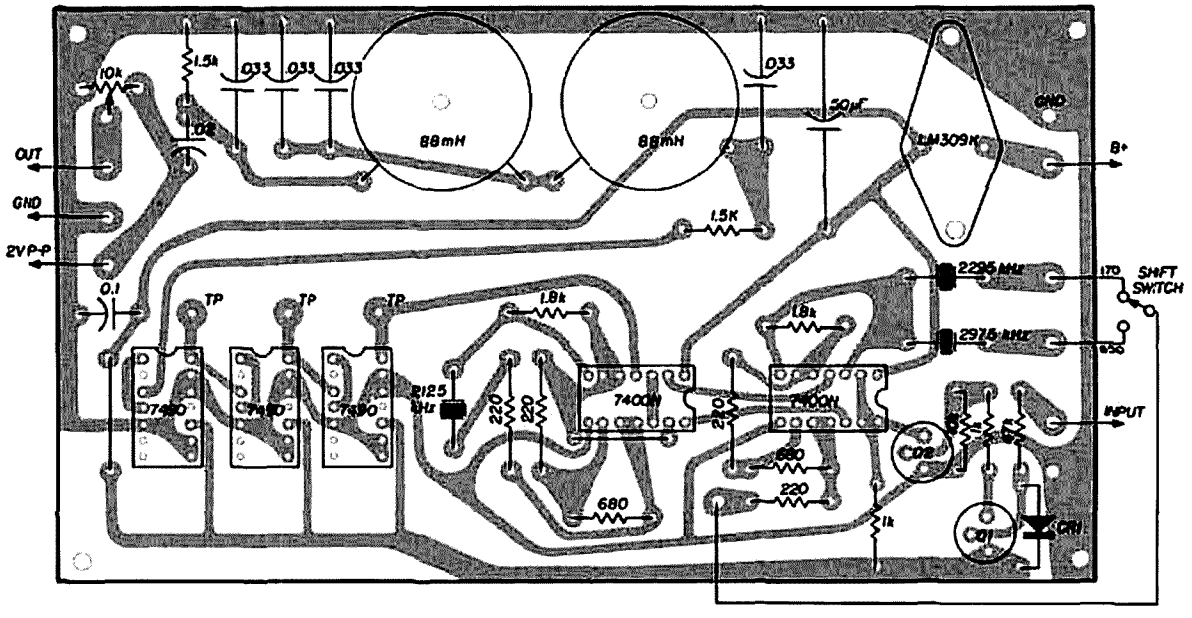


fig. 9. Detailed layout of the AFSK generator circuit board.

weekend project

This audio frequency shift keyer is an example of the type of equipment you can rapidly assemble on your own circuit boards. Though designed for teletype, with slight modifications the same circuit may be used as a crystal calibrator or as a high precision audio oscillator. As a crystal calibrator only one 7400 IC would be needed, a crystal zeroing trimmer added, and a rotary selector switch installed to tap into different points in the 7490 divider chain. Each 7490 IC divides by 5 and then by 2. The divide-by-5 point is at the strap between pins 11 and 14. With a 1000-kHz crystal, 200-, 100-, 20-, 10-, 2-, and 1-kHz outputs would be available. The lowpass filter would not be used.

Perhaps you need a no-drift audio oscillator. Crystals could be selected by plug-in or by rotary switch and different filters switched in as needed. Such an oscillator would not drift even one hertz and would have excellent distortion characteristics. If you want to divide the crystal frequency by some other factor, the 7492 IC will divide by 2, 6 and 12; the 7493 IC will divide by 2, 4, 8 or 16 without changes.

As an AFSK keyer, this unit has excellent performance at low cost. The original design by WB8AAK used RTL



fig. 10. Component mounting points on the paper layout are transferred to the copper-clad board with an awl.

logic. However, the price of TTL devices has dropped so much that it was worthwhile to re-design the circuit to use TTL. The two 7400 gates are 22 cents each and the 7490 decade dividers are only 90 cents each, bringing total logic cost to $3.14. It does not require a counter to set up or adjust.

Even the most inexpensive crystals may be used. JAN stocks the 2125- and 2976-kHz crystals for $1.75 each. Use of a 2976-kHz stock crystal instead of the nominal 2975-kHz crystal only results in a one hertz error in output frequency which is undetectable for most purposes. The remaining crystal – the 2295 kHz unit for 170 shift – may be custom made for $3.00. Everything used in the generator is available from advertisers in *ham radio*. Note that IC supply house prices will be higher than those quoted, which are for surplus devices.

Transistors are another item that may vary considerably in price depending on source. Feel free to substitute. There is nothing critical about the two transistors used in this unit except that they must be silicon types. There are dozens of transistors which may be used. Two types are given as examples.

the circuit

Both crystal oscillators run continuously. Sections of the 7400 ICs gate the oscillator output desired to the 7490 divider string. The output of the 7490 dividers is a square wave which aids in filtering. The TTL dividers are powerful enough to drive the filters and output directly. With the recommended 5 volts on the TTL devices the current drain is 100 mA. The input to the LM309K regulator should be connected to a dc voltage source of from 8 to 30 volts. This may be an existing unit or built for this purpose. The output level is adjustable from -3 dBm down to any value to suit your transmitter.

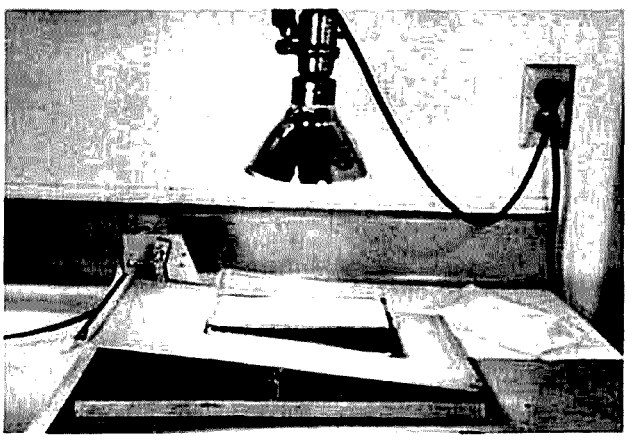


fig. 11. After inking in the circuit traces with the resist pen the board is placed in a tray of etching solution.

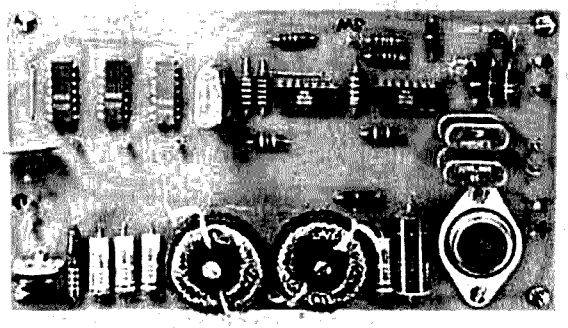


fig. 12. Completed audio-frequency shift keyer board provides crystal-controlled, low-distortion audio tones.

performance

The completed keyer run through checks to verify its performance. The generator was keyed with a 100-Hz square wave and observed on a dual-trace scope to confirm equal *mark* and *space* audio levels and good keying (see fig. 13). Filter ringing accounts for the small transient I observed. Frequency measurement with a digital counter confirmed that output frequencies were accurate to the hertz. Several crystals were tried to confirm this.

Audio distortion was then measured on a Hewlett-Packard distortion analyzer with some interesting results. Harmonic distortion measured only 0.25% at 2975 Hz but 3% at 2295 Hz and 4.2% at 2125 Hz. Scope analysis of the distortion products showed them to be almost entirely third harmonic, as was expected. Obviously, the filter was doing an excellent job at 2975 but only fair at 2125 Hz. The filter design was an old standard and had been used in the original design and other

highly regarded AFSK keyers when 850 shift was popular.

However, 850 shift is now becoming obsolete. If you use only 170 shift, a minor modification of the lowpass filter will give excellent distortion characteristics. Simply add one .022-μF capacitor across each of the .033-μF capacitors in the filter. The distortion will then measure 1% at 2125 Hz *(mark)* and 0.7% at 2295 Hz *(space)* frequencies. This is good enough so that it may be used with any filter-type ssb transmitter.

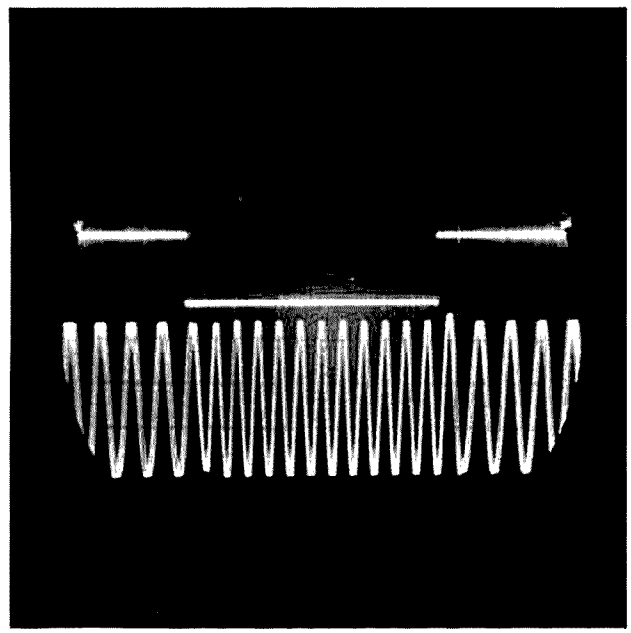


fig. 13. Audio output of AFSK unit provides excellent waveform continuity between mark (2295 Hz) and space (2975 Hz).

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speed standards for international morse code

How to use standard words to determine the words-per-minute speed of your keyer

Gaspard Lizee, VE2ZK, 135 Carillon, Sherbrake, Quebec, Canada

In the formation of the International Morse code used by amateurs, there are two basic *sounds* and three basic *spaces*: the dot, which is 3 time-unit or baud, the dash, which is 3 bauds, a space between each element of a Morse character, which is one baud, a letter space of 3 bauds and a word space of 7 bauds.

Numerals are separated in the same way, and punctuation is preceded by a 3-baud space and followed by a 7-baud space. Therefore, the letter E, one dit, consists of two bauds — a one-baud sound plus a one-baud space. The letter T, one dah followed by a one-baud space, contains four bauds. With this in mind, the ratio of the dit with its space to the dah with its space is 1:2.

counting words per minute

If you key a string of dits for one minute at a constant rate to obtain a total of 25 dits you have used 25 bauds for sounds and 25 bauds for spaces, for a total of 50 bauds in 60 seconds. This rate represents one word per minute. If you only count the number of dits, dividing that total by 25 will also give you the keyer speed in words per minute.

However, it's much easier to count dahs, particularly at higher speeds. Since the ratio of dots to dashes (including spaces) is 1:2, simply count the number of dahs per minute and divide that total by 12.5 to arrive at the equivalent speed in words per minute.

adjusting your keyer

To obtain the proper ratio of dots to dashes first count the number of dits per minute. Then adjust your keyer so you have exactly half the number of dahs in the same time span. When this is done your keyer is properly adjusted, including the spaces. For example, if you key a

string of 250 dits in one minute, you must adjust your keyer to provide 125 dahs in one minute to provide the proper timing ratio.

To arrive at the word-per-minute speed of this keyer setting simply divide the 250 dits by 25 (or the 125 dahs by 12.5) to arrive at 10 words per minute. To calibrate the speed dial for other speeds simply count the number of dahs in one minute and divide by 12.5. If you want to use less than one minute, count the number of dahs in 24 seconds and divide by 5 – it will give you the same result.

automatic keyers

Automatic Morse Keying heads used by the FCC and ARRL are adjusted to a 50-baud word standard. That standard word is PARIS. If the speed of the automatic keyer is set to send the word PARIS ten times in one minute, the sending speed is exactly 10 words per minute.

If you count the total number of bauds in PARIS you come up with a total of 50:

- — — -	P	8 bauds sound 3 bauds element space 3 bauds letter space = 14 bauds
- —	A	4 bauds sound 1 baud element space 3 bauds letter space = 8 bauds
- — -	R	5 bauds sound 2 bauds element space 3 bauds letter space = 10 bauds
- -	I	2 bauds sound 1 baud element space 3 bauds letter space = 6 bauds
- - -	S	3 bauds sound 2 bauds element space 7 bauds word space = 12 bauds
		total 50 bauds

Any other word that has a total of 50 timing units could also be used for the same purpose. It doesn't matter if it has more or less letters so long as it contains a total of 50 bauds.

military stations

Most military stations use a standard 60-baud word to set the speed of their keying heads, using the standard word CODEZ for this purpose. If you count the number of bauds in CODEZ you will come up with a total of 60.

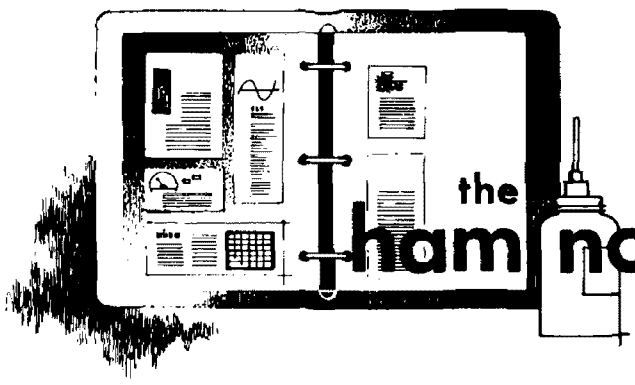
With a 60-baud standard, the number of words per minute appears to be much faster. If you want to experiment a bit, listen to an automatic Morse station that announces its code speed. Copy the station for one minute, count the total number of bauds, and divide by 50. You'll quickly find out what standard they use for setting up their keying heads.

The ARRL code-practice sessions transmitted by W1AW, for example, are based on a 50-baud standard. For comparison, try to copy WAR on a Tuesday evening – it will be much faster than the announced speed. It takes only a minute to figure out that 10 words per minute using the 50-baud standard is only 8.33 words per minute with a 60-baud standard.

The two different standards are a result of two different types of transmission. Most military stations transmit 5-letter code groups; amateur radio operators transmit clear text which uses a larger number of the letters E and I. The two different standards compensate for this difference.

The difference in words per minute with the two standards increases markedly at higher speeds. Sixty words-per-minute clear text (50-baud base) is equivalent to 41.66 words per minute of coded letter groups (60-baud base). This is probably very near the upper possible limit that a human ear is able to differentiate. The current world record, held by the late Ted McElroy, is 75.2 words per minute. This might stimulate someone to suggest a new award, "The Ted McElroy Gold Medal." Anyone interested?

ham radio



5/8-wave whip for 2 meters

A standard 102-inch Citizens Band whip can be easily converted into either a fixed station 5/8-wave groundplane or an effective mobile gain antenna. Gain for either antenna was measured at about 2.5 dB over a quarter-wave groundplane. Both antennas are electrically sound and do not have mechanical joints to corrode and break down. The overall view of the home-station version is shown in **fig. 1**, and the mobile version in **fig. 2**.

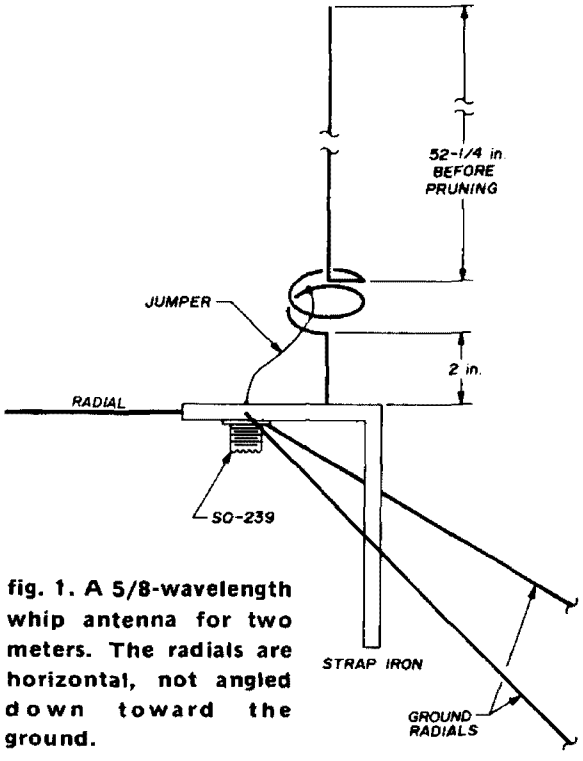


fig. 1. A 5/8-wavelength whip antenna for two meters. The radials are horizontal, not angled down toward the ground.

For the home-station groundplane, measure down 52¼ inches from the antenna's tip and place the antenna in a vise. Using a torch (propane does well), heat the whip cherry red and bend a 90° angle when the metal will allow itself to be bent without undue force. Using a jig made from 2-inch diameter pipe (**fig. 3**), bend the coil slowly while constantly applying heat. Bend 1½ turns with at least ½-inch spacing and finish the coil so that the straight section of the antenna below the coil is lined up with the longer section above the coil. Leave 2 inches of wire below the coil and remove the rest.

The antenna base is made from a seven-inch section of 3/8-inch by 2-inch galvanized flat iron. Drill the flat iron as shown *in* **fig. 4** for an SO-239 coax

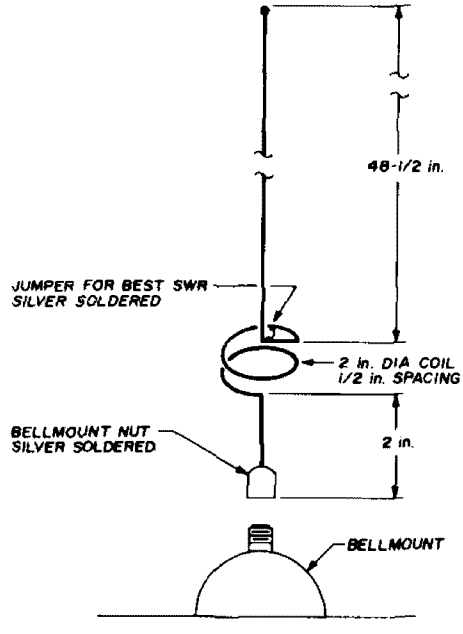


fig. 2. Using the 5/8-wavelength antenna for mobile operation.

connector and its mounting screws. Be sure to ream out the underside of this hole to allow the connector to sit flush with the flat iron. Drill 1/8-inch holes for the ground plane elements and the whip. Heat the iron and make a 90 degree bend as illustrated.

Cut three ground radials from 1/8-inch round rod such as galvanized lathers ceiling rod. Each radial should be 19-inches long. Insert them into the three holes drilled into the side of the bracket and bend the two side radials so there is 120° spacing between all three radials. Insert the whip into its mounting hole until its bottom is flush with the bottom of the hole. Weld or silver-solder the whip and the radials in place. Install the coax connector and solder a short length of solid number 12 copper wire to the center pin (this will go to the tap on the base coil eventually).

The rest of the procedure is simply tuning and pruning. Wrap the free end of the copper wire about one half turn from the bottom of the coil and, with a vswr meter in the feedline, adjust the tap point for best swr. Trim the end of the antenna by filing a notch and breaking off about 1/8-inch of rod at a time. By working back and forth between readjusting the tap and trimming the antenna length, you should be able to tune the antenna for zero reflected power. At this time, silver-solder the copper wire to the coil. Seal around the coax connector and around the copper wire with silicon seal to protect your rf feedline in case of heavy rain or ice build up. It is also worthwhile to paint any spots that might have been exposed to excess heat in the construction. Simply mount the base to your tower or mast.

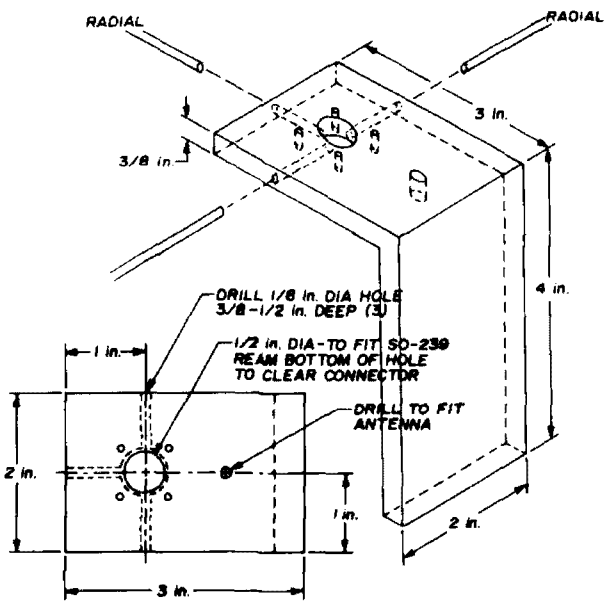


fig. 4. Construction of the antenna base. Horizontal radials are installed and welded in place.

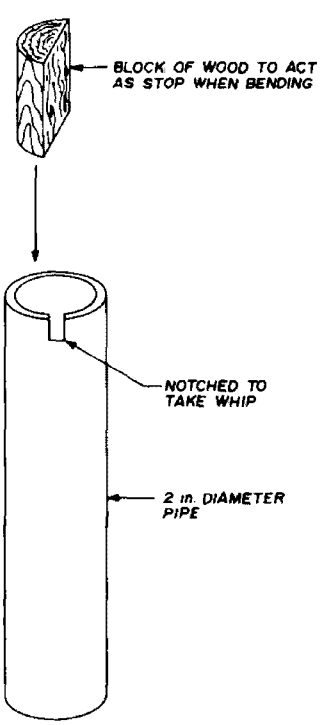
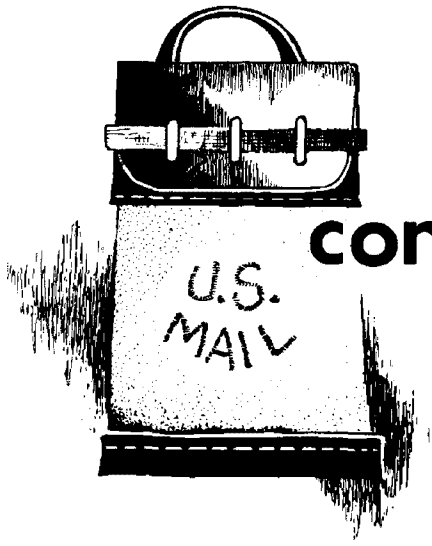


fig. 3. Jig used for bending the base coil (see text).

A mobile version of this antenna is made in basically the same manner. The radiator (after the coil) is cut to 48½-inches rather than 52¼-inches, and the base of the coil is connected to an ordinary bell mount. The base coil is wound just like the home-station version. The mobile antenna is not a grounded system, however. It is tuned for minimum vswr by adjusting the tap point for a short jumper from the junction of the base of the radiator section and the top of the coil to a point on the coil. In my case, by juggling the antenna length and the tap point, I found the best tap point by shorting about 1¼ turns. Silver solder the jumper in place.

John Dobroshinsky, VE3DDD



comments

ssb detector

Dear HR:

Regarding the ssb detector circuit on page 68 of the December, 1972, issue of *ham radio*, the tuning capacitors required to resonate the Collins filter at 455 kHz are omitted, which will result in a substantial insertion loss. The input impedance at pin 1 of the CA-3028 is specified at from 3k to 5k ohms and the impedance at the source of the MPF102 is approximately 2.7k ohms. The lack of matching to these impedances will result in shape and ripple degradation. Therefore, it would be necessary to provide proper matching for either series resonance of the filter which is 600 ohms, or parallel resonance at 100k ohms.

Also omitted is the required bypass capacitor from the drain of the MPF102 to ground.

M.H. Gonsior, W6VFR
Fullerton, California

As I pointed out in my article, there are quite a few things left out. W6VFR has found them.

I do not have any data on the Collins filter and when I breadboarded the circuit back in March, 1972, I felt such a sense of urgency that I did not stop to investigate the impedance of the filter. Now that I have been given the impedance values, finding the capacitor values will be easier.

The gain of the amplifier was marginal; however, I was testing the detector, not the amplifier. It is too bad that I have already disassembled the circuit. It would be interesting to see how much of an improvement could be gained by matching the impedances.

In the circuit, the MPF102 and the NE561B were physically close together and the 0.5 µF capacitor shown near the IC served as a bypass capacitor for both devices.

Max Robinson, K4ODS

touch-tone modification

Dear HR:

I have not had a single transistor failure due to the mobile touch-tone circuit as originally presented in the August, 1972 issue of *ham radio*. However, I will admit that WB8NAT's change was better.

The following addition to the unit, one which I did not think of originally, will prevent any filter capacitors from charging up, discharging, then re-charging after the +12 volts dc has been removed. In short, the switch labeled *burst/pad* for dc switching is changed to a dpdt switch rather than a center-off spdt unit. The improved circuit is shown in fig. 1.

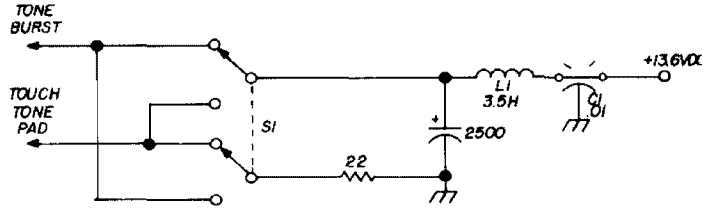


fig. 1. Improved Touch-Tone circuit.

Bill Lambing, WØLPQ
Marion, Iowa

satellite communications

Dear HR:

In the satellite communications article by K1TMA in the November, 1972, issue the comment concerning antenna gain on page 55 states: "The primary concern of the system loss formula is to know the amount of rf gain there is in a particular array in the desired direction of transmission, relative to half-wavelength dipole or that hypothetical antenna, the isotropic source."

Actually, there is a 2.14 dB theoretical gain that a dipole has over an isotropic radiator, so to be on the level in calculations the isotropic radiator should always be chosen as the reference antenna. This will entail a knowledge of the gain-over-isotropic of a particular array. It should be borne in mind that some handbooks use gain over isotropic, while others, notably the ARRL *VHF Manual*, use gain over an ideal dipole. Most commercial manufacturers use gain over isotropic in the specification sheets. In any case, use of the system loss formula in engineering fashion means referencing all antennas to the point source or isotrope.

Another point not specified is the reference for computing transmitter output power in dB; power output should be specified in dBW, or dB above 1 watt. While this parameter is not specified in the system loss formula, it is nonetheless important; the result of the system loss formula, as indicated in the article, will be a large negative number in dB. Thus,

$$LS + \frac{S+N}{N} = P_O + S$$

where LS is system loss, P_O is power output in dBW, S is effective receiver sensitivity in dBm, and (S + N)/N is the required signal to marginal communications (use 3 dB for CW, 6 dB for ssb).

P.H. Bock, Jr. K4MSG
Avon, North Carolina

sstv synch generator

Dear HR:

I have just finished reading WA2EWO's article on the sstv synch generator, and in my opinion, the design is too costly and complex. I have a similar unit which is built on a piece of Vector board, 2 x 5 inches. Total cost for all the ICs is $4.04 including three 74121 single-shots, one 7473 divide-by-four, one 7490 divide-by-ten and one 7492 divide-by-12 connected as shown in fig. 2. The cost of WA2EWO's synch generator is nearly $20.

Gordon P. Stanys, W1IA
Stamford, Connecticut

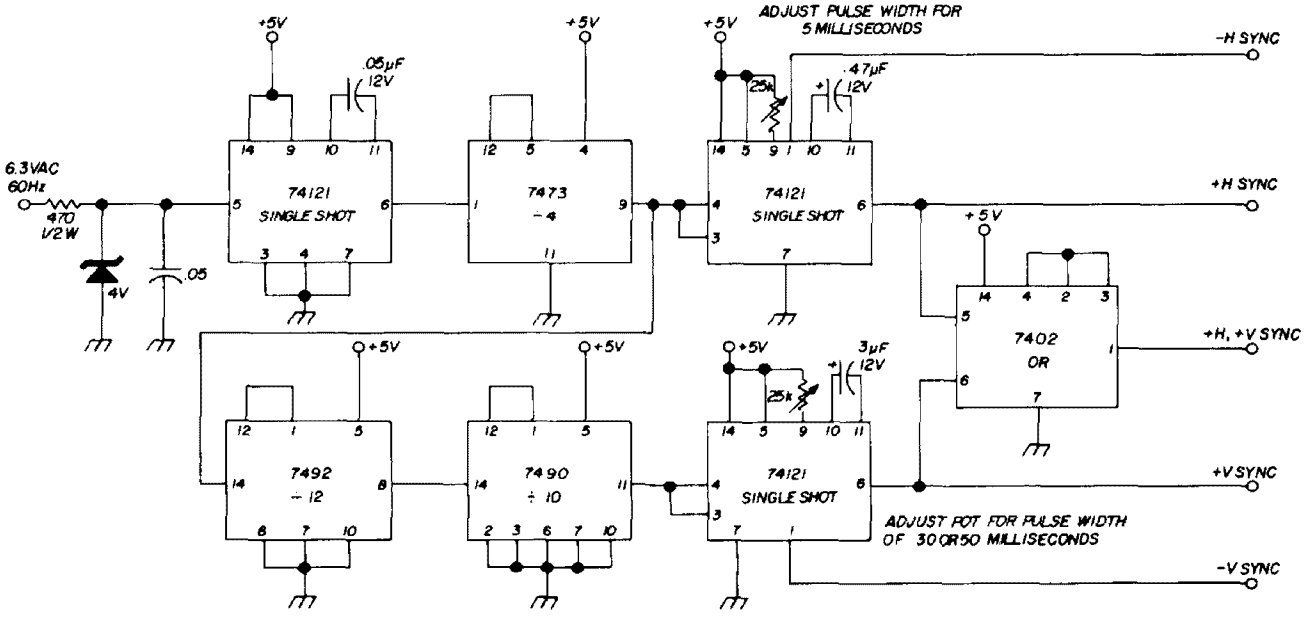
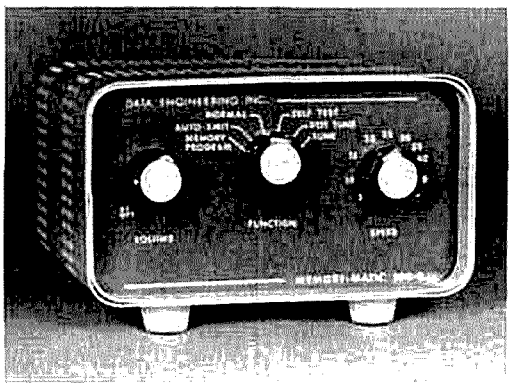


fig. 2. Block diagram of the sstv synch generator.

new products

memory-matic 500-B

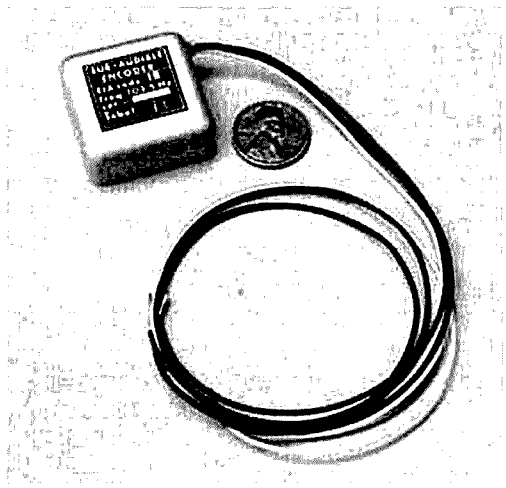


Data Engineering's reprogrammable memory keyer has been improved and updated. The Memory-Matic 500-B today features a choice of 500-bit or 800-bit memory and provision for remote control of the message start and message stop function. The dot to dash weight ratio is also completely variable and the type of keying is selectable — choosing, dot memory only, dash memory only, no memory operation or full dot and dash memory operation — all with or without automatic character and word spacing.

A memory stop switch allows the insertion of some information in the midst of a list of sequential memory transmissions. An example of this operation would be the manual insertion of a station's signal report in a preprogrammed message including a greeting, your location and your name.

The unit features iambic squeeze keying, built-in adjustable monitor and speaker, rf-proofing, and built-in 117 Vac power supply. The new keyer sells for $198.50 with a 500-bit memory reprogrammable memory or $219.50 with an 800-bit reprogrammable memory. Full details on all the features of this new keyer are available from Data Engineering, Inc., Ravensworth Industrial Park, 5554 Port Royal Road, Springfield, Virginia 22151 or by using *check-off* on page 110.

sub-audible encoder



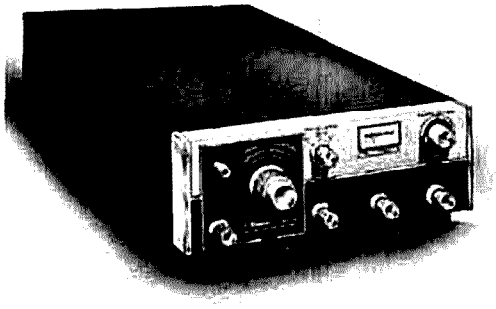
The new Tobel CTCSS encoders fill the void for a low cost, sub-audible, miniature tone encoder and are ideally suited for mounting in solid-state transceivers, as well as, larger tube-type 2-way radios. Extensive field tests have proven the CTCSS encoders to be superior in both design and performance to older mechanical reed-type devices. Tobel encoders provide faster starting time, lower power consumption and immunity to damage from mechanical vibration. Low power consumption (2.5 mA at 12 V) and small size make Tobel encoders a natural for small hand-held transceivers and other battery operated communications equipment. Minimal effort and expense is required to convert your FM radio to sub-audible tone coding.

CTCSS (Continuous-Tone, Controlled Squelch System) makes possible positive repeater access and receiver squelch control under conditions of high noise, weak signals, and mobile flutter. CTCSS encoders are available for any EIA code L1 thru 7A as standard units and must be specified when ordering. Non-standard frequencies are available on special order.

More Details? CHECK–OFF Page 110

For more information, write to Tobel Electronics, 7920 Alida Street, LaMesa, California 92041, or use *check-off* on page 110.

solid-state high-frequency ssb transceiver



Swan Electronics has just introduced a full line of completely solid-state ssb transceivers for use on 80 through 10 meters. Models are available for 200, 100 and 15 watts PEP ssb, and all transceivers feature all solid-state, broadband transmitting circuits, as well as full coverage, 80 through 10 meters, CW and ssb. The new Swan solid-state transceivers have built-in VOX and a built-in noise blanker with variable blanking control. The transmitter circuits, which require no tuning, have infinite vswr protection from an open to a short circuit, so it's impossible to damage the final power transistor because of poor antenna matching.

Receiver sensitivity for the solid-state Swan transceivers is less than 0.5 microvolt at 50 ohms impedance for a signal-plus-noise-to-noise ratio of 10 dB. Minimum image rejection is -55 dB at 30 MHz, increasing to better than -75 dB at 3 MHz. Audio response is essentially flat from 300 to 3000 Hz ±3 dB. Audio output, to an external 3.2-ohm speaker, is 4 watts with less than 10% distortion. I-f selectivity is provided by a 2.7-kHz wide crystal filter with a 1.7 shape factor.

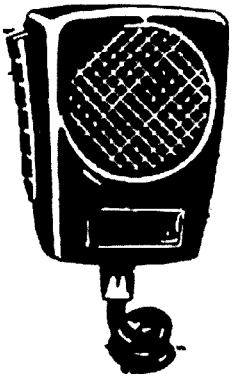
In the transmit mode the unwanted sideband is down more than 50 dB and the carrier is suppressed greater than 60 dB. Distortion products are approximately -30 dB. Transmitter output impedance is 50-ohms nominal. An audio sidetone is

Hi Hobby Industry

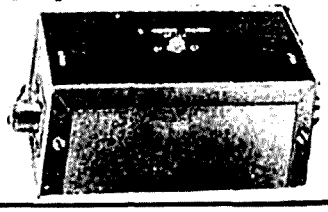
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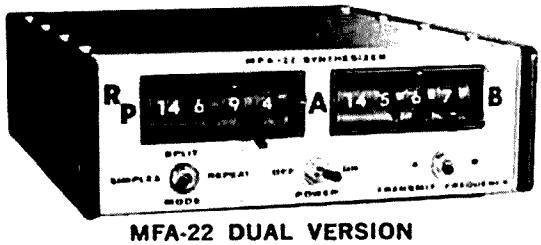
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provided to CW monitoring, and CW keying is equivalent to grid-block keying and features semi-break-in with VOX.

The vswr protection circuit cuts output power by 20% when the vswr is 3:1 or higher. Under these conditions, infinite vswr protection is provided, including open or short circuits.

Three models are available, the SS-200, with 200 watts PEP ssb minimum on all bands and 200 watts dc input on CW; the model SS-100, 100 watts PEP minimum on ssb and 100 watts CW; the model SS-15, 15 watts PEP minimum on ssb and 15 watts CW. All transceivers are designed for 13.5 Vdc power supplies, the nominal supply voltage of a 12-volt automobile system. All units require 500 mA in the receive mode. For ssb transmission, the SS-200 requires 6 amps average; the SS-100, 3.5 amps average, and the SS-15, 800 mA average. For CW operation, the current drain is higher, being 20, 11 and 2 amps respectively for the SS-200, SS-100 and SS-15. Accessory power supplies are available for operation in the home station from 117 Vac utility lines.

Other accessories include the SS-208 external vfo for split-frequency operation, the 61OX crystal-controlled oscillator, the SS-16B 16-pole super-selective crystal filter with a 1.28 shape factor, and the SS-1200 1200-watt PEP linear amplifier using four 6LQ6 tubes for 117-Vac operation.

The SS-15 ssb transceiver is priced at $579.00. The SS-100 is $699.00, while the SS-200 is $779.00. The 1200-watt SS-1200 linear amplifier is priced at $299.00.

For more information, write to Swan Electronics, 305 Airport Road, Oceanside, California 92054, or use *check-off* on page 110.

great circle charts

As a service to radio amateurs WB5CBC is making available, to amateurs anywhere in the world, a comprehensive chart of great-circle bearings to 660 locations throughout the world for the unbelievable price of $1.00, postpaid. These

More Details? CHECK—OFF Page 110

charts, which are a handy 8½ by 11 inches, contain great circle bearings to the nearest degree, distances in miles and kilometers, and return bearings. Directional accuracy of the charts is to the nearest degree while distances are to the nearest mile and kilometer.

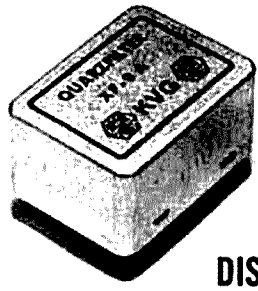
The magnetic bearing is not given on these charts because magnetic declination at any location varies from year to year, sometimes as much as 10°. It is easier to zero your antenna on true north (using the North star) and forget it.

The return bearing, a particularly useful feature of this chart, is the bearing from the distant location to your station. The return bearing is *not* a simple 180° difference from the outward bearing, but must be derived through spherical trigonometry. The ability to give a station being worked the correct bearing for aligning his antenna will produce results which are nothing short of amazing. When the chart indicates a distance greater than 8000 to 10,000 miles, the long path, which is simply 180° opposite the indicated bearing, can be considered. The same is true for the return bearing.

Charts consist of six pages and are available for any location in the world. To order a chart, the following information is needed:

1. Name and postal address.
2. The city for which the chart is wanted (include state and/or country).
3. If the population of the city is less than 10,000 or if the location is a rural area, also include the latitude (indicate north or south) and longitude (east or west) in degrees and minutes.
4. Include payment of $1.00 (postpaid worldwide) or $1.75 (airmail worldwide). Cost to overseas stations is 7 IRCs (12 IRCs for airmail).

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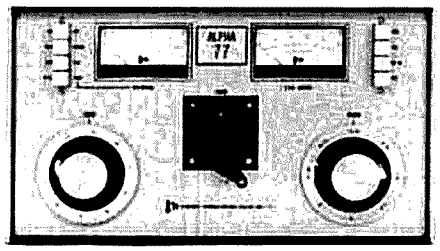
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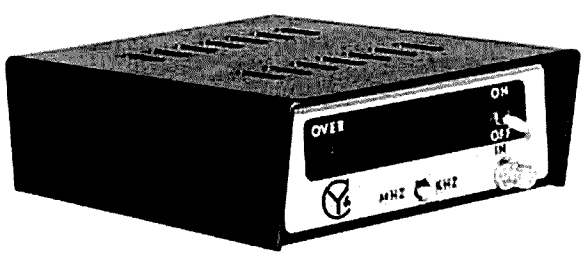
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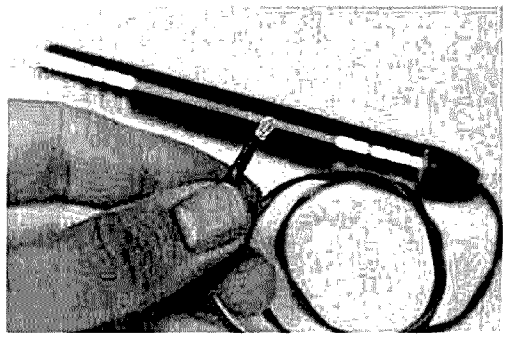


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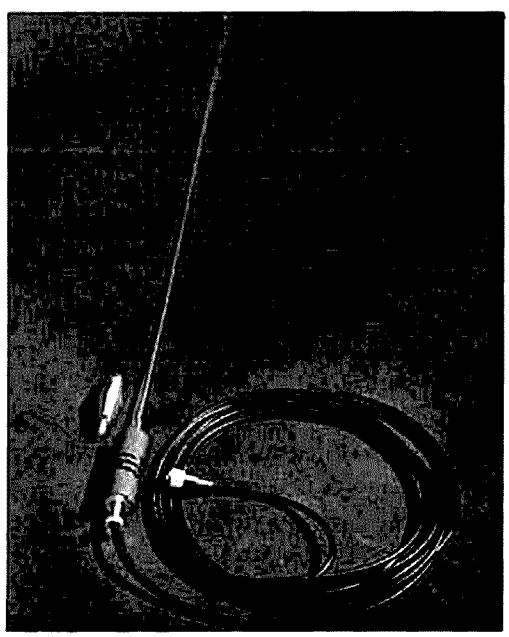
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The price for the inspection lamp and flashlight is $2.50 postpaid. Write to Spacetrom, Box 84, Broadview, Illinois 60153, or use *check-off* on page 110.

mobile gain antenna



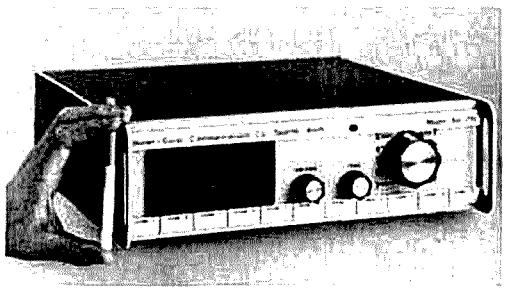
The new Antenna Engineering M-series antenna is a full 5/8-wavelength vertical whip with a bottom-load matching transformer which is designed for vehicle mounting. The adjustable threaded bushing allows for either roof-mounting or mounting through double-panels. The

More Details? CHECK-OFF Page 110

matching transformer is encased in a fiberglass sheath which is nearly indestructable. The whip is of 1/8-inch spring-temper type 302 stainless steel; the coil tip unscrews to accept a chrome-plated spring (optional at extra cost) for severe service applications.

This antenna is available for all amateur and commercial frequencies in the 140-175 MHz, 220-225 MHz and 420-470 MHz bands. The M-series antennas are at dc ground for dissapation of static, and are supplied with 20 feet of type RG-58A/U coax and ugf connector. Prices range from $16.95 for the 140-175 MHz version to $15.95 for the 420-470 MHz version. For more information, write to Antenna Engineering Company, Inc., Box 19449, Indianapolis, Indiana 46219, or use *check-off* on page 110.

solid-state ssb radiotelephone

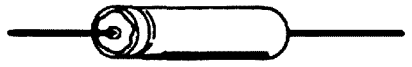


Stoner-Goral Communications has introduced their new ssb radiotelephone. It is believed to be the first all solid-state 100-watt MF/HF marine single-sideband radiotelephone and was designed as a direct replacement for obsolete a-m equipment.

The unit operates on frequencies between 1.6 and 9.0 MHz and has a capacity for 11 duplex or 22 simplex channels. Any channel may be programmed for AME as well as suppressed or reduced (-16 dB) carrier operation. A variety of power supplies and accessories permit its use in land-based, mobile and marine stations.

The exciter portion of the radiotelephone is an up-conversion scheme and uses integrated circuits almost exclusively. Simplified channelizing involves inserting crystals and "netting" to fre-

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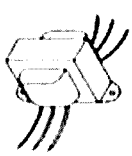
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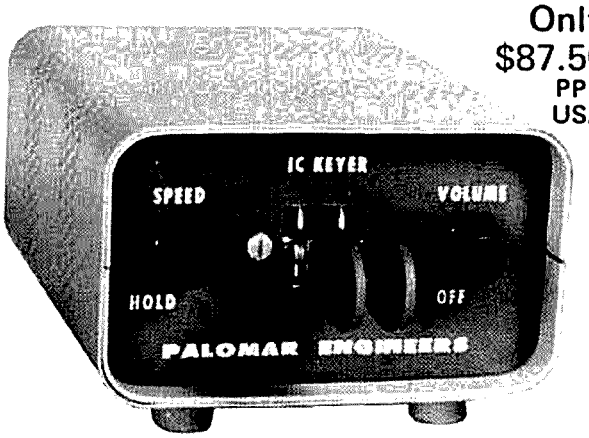
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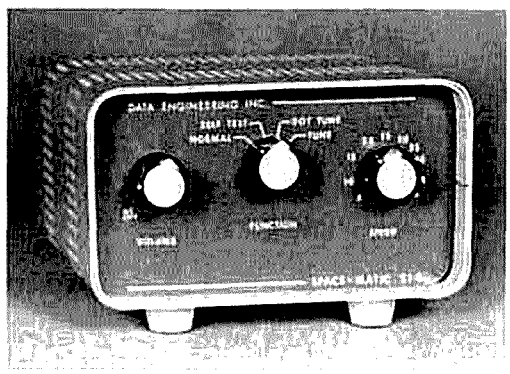
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For additional information, contact Donald L. Stoner, Director of Marketing, Stoner-Goral Communications Co., Box 233, Mercer Island, Washington 98040, or use *check-off* on page 110.

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More Details? CHECK–OFF Page 110

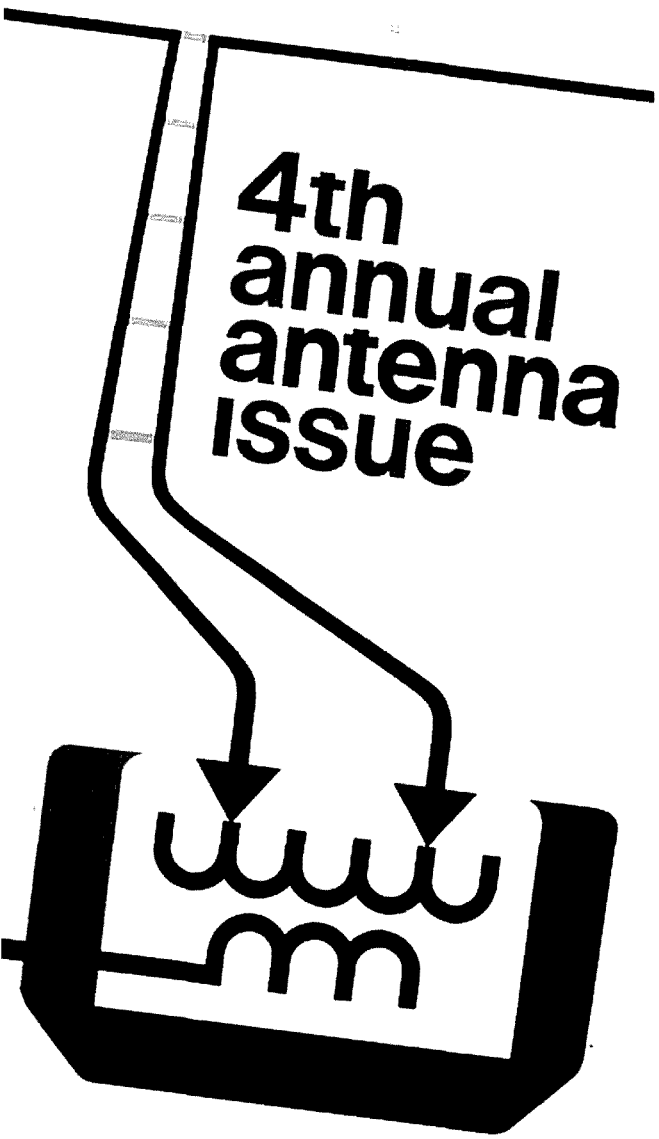
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ham radio magazine

MAY 1973



4th annual antenna issue

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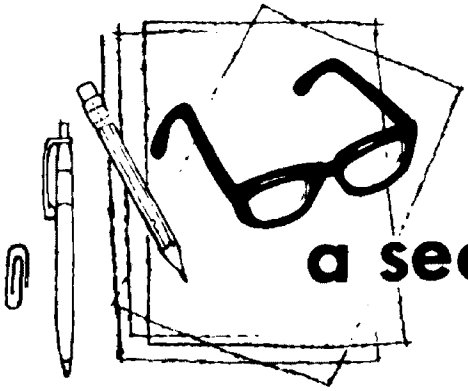
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a second look

by jim fisk

One of the more serious problems which has plagued amateurs who operate on the vhf and uhf bands is that of rf interference — interference with television sets, stereo and hi-fi systems, intercoms and even tape recorders, practically any consumer electronics equipment that uses solid-state circuitry. Nor is the condition restricted to vhf. I was involved in a tense situation with one of my neighbors a few years ago when my 20-meter ssb signal wiped out a recording on his expensive solid-state tape recorder. No matter of electronic tricks, other than a big copper shield box, completely eliminated the interference, and the problem was only finally resolved when he and his damn recorder departed for Arizona!

Amateurs are not alone in their plight, either. Television, am and fm broadcast engineers are faced with daily interference complaints from nearby homeowners who don't particularly appreciate the cacophonic symphony of football telecast interference to "Saturday Afternoon at the Opera," or vice versa.

What is needed is Congressional or FCC action to require all manufacturers of tv sets, stereos and am receivers to build interference suppression into their designs. Some lead bypassing and narrow filters at the input would go a long way toward solving the present problem.

Last year, Congressman Charles M. Teague of California introduced a bill in Congress which would amend the Communications Act of 1934, requiring that "apparatus designed to receive broadcasts" would meet FCC standards to be adopted so that "all interference from any amateur station operating on its assigned frequency will be filtered out." Unfortunately, no action was taken by Congress before it adjourned.

However, Mr. Teague has re-introduced his bill to the 93rd Congress (HR 3516) so we may have some action this year, but getting the bill re-introduced is a small part of the battle. First, the Chariman of the House Interstate and Foreign Commerce Committee must be convinced to call a hearing on the bill. To do that we will need a lot of letters. Once approved by the Committee, the bill can be presented to the members of Congress for a vote. However, all this will require a massive letter-writing campaign. You can bet that the manufacturers will be lobbying against any legislation that will make their product more complex or more expensive, so amateurs will have to work extra hard to gain favorable and speedy Congressional action.

Right now, while you're thinking about it, write a letter to Mr. Staggers, Chairman of the House Interstate and Foreign Commerce Committee, as well as to your own Congressman, indicating your support for HR 3516. Letters of support should also be sent to Mr. MacDonald, Chairman of the Communications and Power Sub-Committee; in all probability it is this Sub-Committee which will hear HR 3516. The addresses are given below.

The Honorable Harley O. Staggers
Chairman, House Interstate and
Foreign Commerce Committee
2366 Rayburn Building
Washington, D.C. 20515

The Honorable Torbert H. MacDonald
Communications and Power
Sub-Committee
Room 2125 Rayburn Building
Washington, D.C. 20515

Because of the importance of this bill to amateurs, I will make petitions available to interested readers who send me a self-addressed, stamped envelope. This petition can be circulated among your friends and neighbors. We need all the help we can get.

Jim Fisk, W1DTY
editor

low-cost RX impedance bridge

Complete calibration and application information on the simple W2CTK impedance bridge

Bill Wildenhein, W8YFB, 41230 Butternut Ridge, Elyria, Ohio 44035

In the September, 1970, issue of *ham radio*, W2CTK described a simple impedance bridge capable of independent readout of R and X values.[1] Readers familiar with laboratory bridges will appreciate the potential of this instrument, but may doubt the accuracy of such a simple device. Less experienced readers may have shied away from this bridge since the original article showed only the use of Smith charts to interpret the bridge readings.

This article is written to compare the accuracy of W2CTK's bridge with a Boonton 250A. Some construction points are clarified, techniques are presented to substantially improve measurement accuracy, a wider range of applications is presented, and a step-by-step procedure is included for those who are unfamiliar with bridge calculations.

construction

First, let's tackle construction: Fig. 1 shows the complete schematic of the impedance bridge, but only a part of the schematic will affect accuracy. Fig. 3 shows the basic bridge. This portion is the heart of the unit and should be wired with short, heavy leads. A suggested layout is shown in fig. 4. The photograph may help with parts placement. Notice that the rear of the panel around the basic bridge has been stripped of paint to insure good grounding of bridge components. Be sure to use a *composition*-type potentiometer; a wirewound pot is completely useless.

A pot with a linear taper will result in the most useful dial calibration. The 56-ohm resistor can be any value from 47 to 68 ohms with no affect on final accuracy, once the R dial is calibrated with that particular resistor. Of course, this resistor must also be a composition type. Some wirewound resistors look like composition types, but are identified with one extra wide color band. Some, made during the Great War and sold surplus for years afterward are, in fact, wirewound, but do *not* have the identifying color band.

The variable capacitor can be any value from 150 to 365 pF. The lower values will limit the X range of the bridge, but have slightly better dial readability. The upper ranges have the opposite characteristics. Any of the values will result in an accurate bridge, but a 250-pF capacitor is about optimum. The diode can be any germanium type, similar to a 1N34. The .005-μF disc capacitor and the diode should have very short leads. Heat sink the diode while soldering it in place.

calibration

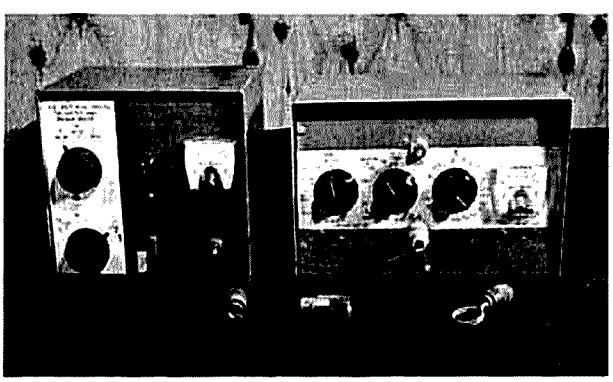
For convenience I will go over the calibration procedure. First, connect the basic bridge as shown in fig. 5. This hookup allows calibration of the R pot

with dc. W2CTK is correct in his statement that rf measurement accuracy will not be impaired with this procedure. The rather shallow null obtained with this bridge is the major source of error, and will limit you to about 5% of calibration accuracy on both R and X measurements. Total error consists of this 5% *plus* the percentage error of resistors and capacitors used to calibrate the bridge.

For convenience, the battery can be fed into the RF IN jack. The 10k resistor is included so you don't bang the meter hard if the bridge is badly unbalanced. Connect a known resistor across the LOAD jack. Lead length is of no consequence here, since this is a dc calibration. For the same reason, wirewound precision resistors are also permissible. Touch the end of the 10k resistor to the center pin of the RF IN jack. Adjust the 100-ohm pot to bring the meter down to zero. Near zero you can touch the battery end of the 10k resistor to the center pin to improve the sensitivity of the adjustment.

If you continue turning the pot in the same direction, it will go below zero. The correct point is where the meter needle just reaches its zero resting position. Mark the R dial with the value of resistance you used for reference. Repeat the procedure for various resistors, or combinations of resistors, to obtain as complete a calibration as you wish. If you use new 5% resistors, your probable accuracy of R readout will be 10%.

Best R range for this bridge is approximately 20-ohms to about 3000 ohms.



Homemade RX impedance bridge, right, and rf source, left.

Below 20 ohms you may be limited by available X dial range – even with an essentially resistive load. Above 3000 ohms the dial graduations become quite close together, limiting the accuracy of readout. My bridge is calibrated to 5000 ohms, but it is difficult to get precise readings at that point. If you wonder about the variable capacitor in the circuit during this part of the calibration, forget it. It merely forms a convenient connection to the load jack.

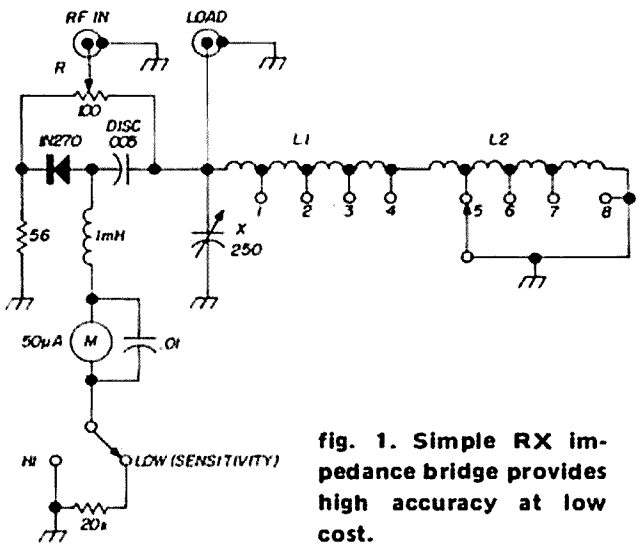


fig. 1. Simple RX impedance bridge provides high accuracy at low cost.

After R dial calibration is complete, disconnect the lead from the variable capacitor to the 100-ohm pot. The variable capacitor is now the only component connected to the LOAD jack. The easiest way to calibrate the capacitor is to connect a capacitor bridge to the load jack and calibrate the capacitor in about 10-pF increments from 250 to 100 pF, and in 5-pF increments below 100 pF. There is no need to be concerned about stray capacitance here, so long as it is minimized. You are interested primarily in the difference between two readings, not so much the absolute value.

If you have no capacitor bridge available, you can get an excellent calibration using the external hookup shown in fig. 6. A coil is connected to the load jack with provision for placing reference capacitors across the LOAD jack. A source of rf power is link coupled to this coil, and a vtvm ac probe is connected across the coil

to read the peak voltage when the coil and X dial capacitor is resonant at the rf input frequency.

The important thing here is to make a rigid setup, mechanically. The coil and coupling link must be quite rigidly held with respect to one another. The rf source must be stable. The entire setup must be held in a fixed position relative

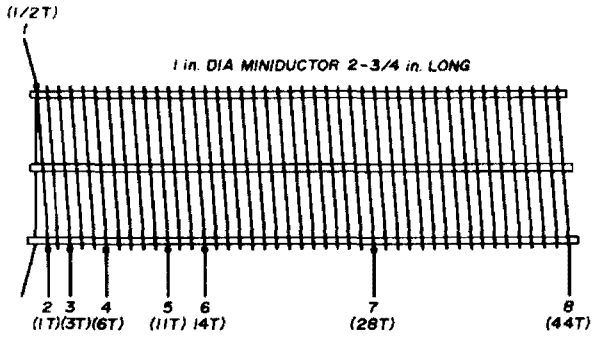


fig. 2. Construction of L1 and L2 in the RX impedance bridge.

to the metalwork of the impedance meter to minimize variations in distributed capacitance.

The coil can be a 3-inch length of 1-inch diameter 16-pitch Miniductor. Later, this same material can be used for the bandswitched coil, so there is no loss. Cut the coil wire about 3 turns in from one end. Heat the ends and push them out of the support bars to provide connecting leads an inch or so long. Do not cut the plastic support bars between the two coil segments. They will help maintain the coupling. Mount the coil as shown in **fig. 6**.

With the setup of **fig. 6** assembled on a scrap of wood and connected to the LOAD jack of the impedance meter, first set the 250-pF X dial capacitor to maximum capacitance. Mark the dial accurately to indicate maximum. (It is advised to do the same for the R pot. This gives you a reference point in case the dial setscrews loosen and it is necessary to reposition them without re-calibrating.)

Feed an 80-meter rf source to the link and adjust the coil tap for a peak reading with the X dial capacitor at maximum. When this point is found, it must be maintained for the balance of the capacitor calibration. Solder it carefully in place. Connect, say, a known 100-pF capacitor to the binding posts. Re-peak the vtvm with the X dial capacitor. That point is 250 minus 100 pF = 150 pF. With only two 100-pF, one 50-pF and one 25-pF mica capacitor you can calibrate a 250-pF bridge capacitor in increments of 25 pF throughout its range. With care, you can estimate the intervening 5- or 10-pF increments with acceptable accuracy.

It would be worthwhile to buy this group of capacitors new in 5% tolerance or better. (These are available in silver mica to 1% quite reasonably.) If you use a grid dipper for this calibration you will find the dipper frequency pulls as the X dial approaches resonance, making it difficult to maintain exact frequency. For that reason it is better to use an 80-meter crystal-controlled oscillator.

This completes the calibration. You can now complete the wiring of the bridge. Fig. 2 shows a suggested coil made from the Miniductor coil stock. If you used other than the suggested 250-pF variable capacitor the tap points will have to be changed. The only criterion for selecting tap points is this: On each amateur band from 3.5 to 30 MHz set the taps so that the band can be tuned at near maximum (about 200 pF) and near minimum (about 50 pF). This is necessary to get best X dial range in measurements. The easiest way is to unsolder one end of the wire from the 100-ohm pot to the variable capacitor, install the coil, then set each tap using a grid dipper. (The 100-ohm pot would swamp the tuned

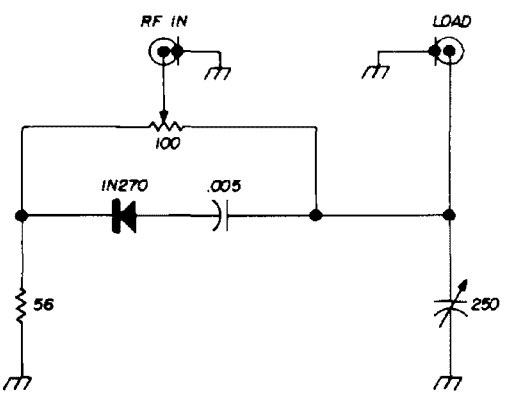


fig. 3. Basic circuit of the impedance bridge.

circuit if left in, and would make it difficult to find a dip.) Unsoldering and re-soldering this lead will not affect bridge calibration if the wire is carefully repositioned.

rf source

The next concern is a source of rf to drive the bridge. Most solid-state dippers will lack the power necessary to do a good job. Some vacuum-tube grid dippers will also be marginal. In any case, a dipper has poor stability, so I would advise building an all-band crystal oscillator. The one shown in fig. 7 covers 3.3 to 33 MHz with fundamental FT-243 surplus crystals. It will operate on the fundamental, second, third or fourth harmonic and provide ample drive for the impedance bridge. In my case I use a pair of 80-meter crystals and a pair of 40-meter crystals to do 90% of my work.

Layout is not critical. Just keep your rf leads short and direct, and your bypass capacitors installed with short leads. It is worthwhile to use up some of the old vacuum-tube gear in the junkbox on projects like this. Take an evening to do the metalwork nicely! It can be housed in a wrap-around made from a discarded piece of galvanized iron flashing. A spray can of gray lacquer will give it a professional touch. This oscillator will find many other bench uses other than as a bridge driver. For example, it can be used as a driver for experimental solid-state power amplifiers, as a temporary local oscillator for converters or transmitter mixers, as a frequency marker, as a QRP CW transmitter, etc. I find it useful as an rf source for calibrating wavemeters for novices and as a reference antenna driver in antenna tinkering. If you have an old set of plug-in coils, they can be used with somewhat improved efficiency.

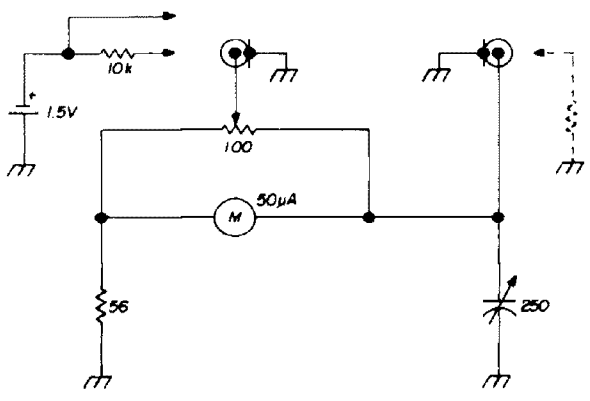


fig. 5. Test setup for calibrating the R potentiometer.

bridge operation

Now, let's put the bridge to work. The following abbreviated procedure will be followed with some notes on methods of improving accuracy and interpretation of the results.

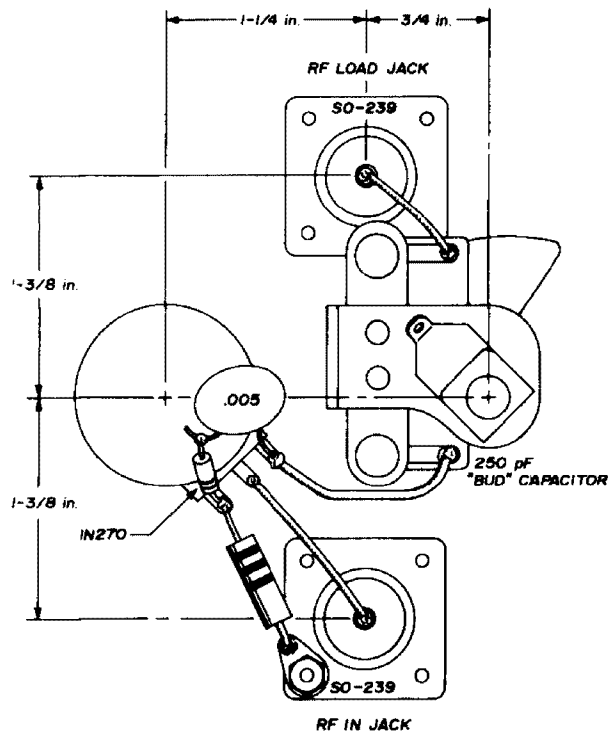


fig. 4. Suggested layout of the bridge resistor and capacitor.

initial balance

1. Set the R dial to maximum resistance.

2. Feed rf at the desired frequency into the RF IN jack.

3. With nothing connected to the LOAD jack, alternately vary the R and X dials for best null.

4. Note the exact reading in pF on the X dial.

measurement

5. Connect the unknown load to the LOAD jack.

6. Again adjust the R and X dials for best null.

7. Note the exact readings of the R and X dials.

8. The R dial reading obtained in step 7 is

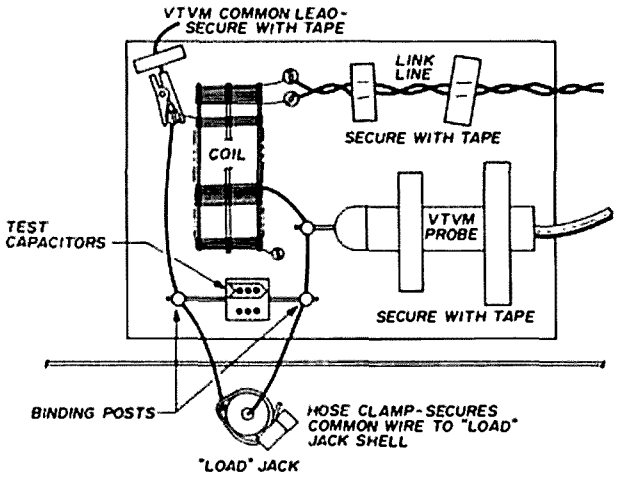


fig. 6. Test setup for calibrating the X capacitor.

the actual resistive component of the load. The *difference* between the X dial reading of step 4 and the X dial reading of step 7 is the reactive component of the load (C_p).

9. If both X dial readings are the same the load is purely resistive.

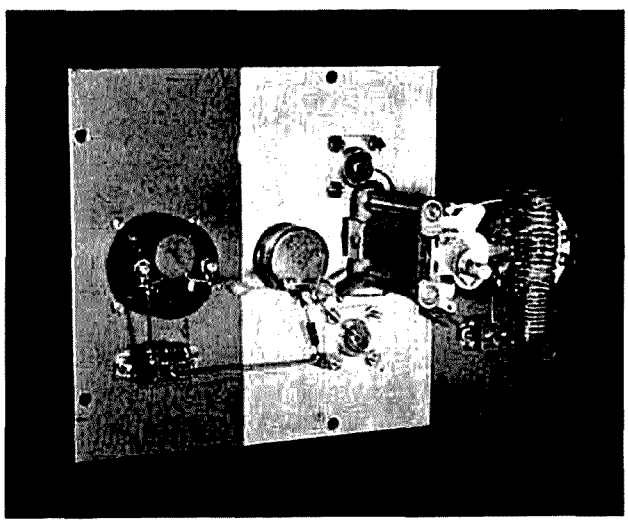
10. If the X dial reading of step 7 is a higher capacitance than in step 4, the load is *inductive*. If the step 7 reading is lower, the load is *capacitive*.

A little trick I use to get the best null is to view the meter from an angle so a little slit of white background shows through between the meter needle and the zero mark on the meter. In this way very small meter variations can be seen. You will find that by using a crystal-controlled oscillator for the bridge driver the X dial reading in step 4 will be very accurately repeated, so once you know that number of pF for a given amateur band on a given bandswitch position, you can use the same crystal and skip the initial balance procedure. This is a big help when many measurements are necessary.

For example, I had an experimental tilted longwire antenna I wished to prune for best compromise on a number of bands. Since each pruning required a walk down the valley to the willow tree and back, it could be a full afternoon of work. However, knowing accurately what the X dial would be for initial balance setting it was possible to make a fast measurement on all bands from 80 through 10 meters and really see where the antenna was resonant. Inside of an hour I had it accurately set.

If, in step 6, you find it impossible to obtain a null, it means you ran out of X dial range. For example, if the X reading obtained in step 4 were 25-pF and your load was *capacitive* by 50 pF obviously you would run out of dial range. So, switch to another bandswitch position, repeat the *initial balance,* and perhaps now the reading in step 4 will be 200 pF. Again bridge the unkown load, and this time (assuming the load was 50 pF capacitive) the X dial will read 50 pF lower than the initial balance setting and you will have no trouble finding a null.

Of course, if the load is horribly reactive, it may be impossible to find any null. You could then grid dip the load to try to find where it was resonant, and this will tell you which way to go to get the load into the bridge range. If the load cannot be easily grid-dipped, you can



Construction of the RX impedance bridge.

make bridge measurements at other frequencies to find a frequency where it does fall within the bridge range. This same problem is common in laboratory impedance bridges. The Boonton 250A doesn't have nearly as much X dial range as W2CTK's simple bridge.

homebrew bridge and calculated it will look like 21.35 ohms. Try to approach that kind of accuracy with an *antennascope* or other simple impedance bridge.

However, I felt that, although the difference in impedance measurement was negligible, many experimenters might

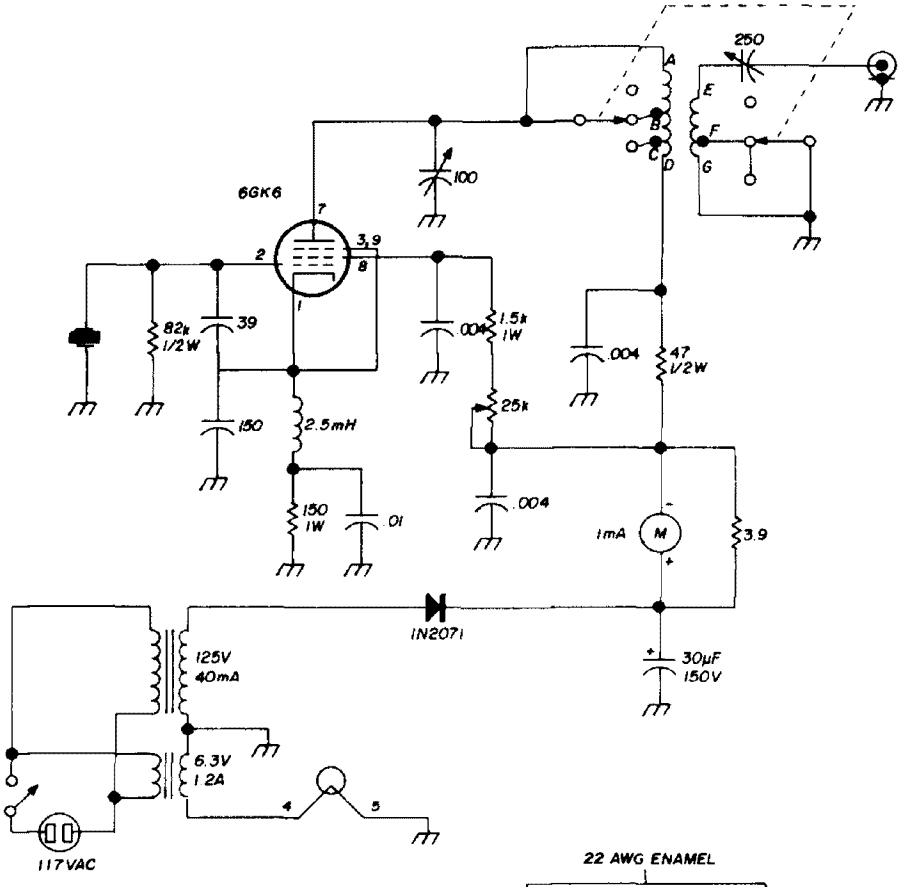
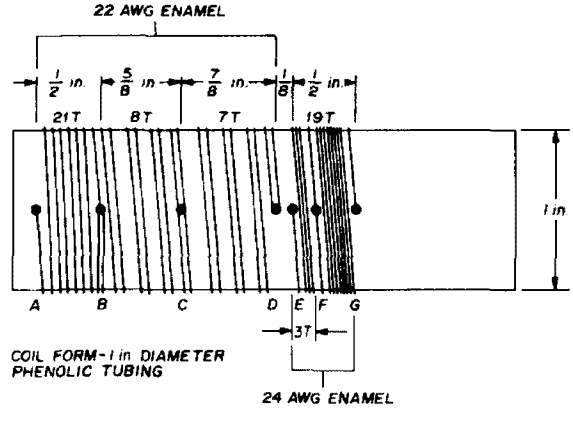


fig. 7. Circuit for the 3.5- to 33-MHz oscillator/multiplier used as an rf source with the RX impedance bridge.



From a point approximately 150 or 200 ohms and higher on the R dial the foregoing procedure will yield excellent accuracy. Below about 150 ohms an error creeps in which will add approximately 5% additional error, at worst. The X dial does not follow a Boonton 250A. At first glance the error appears to be large, but it is not. For example, a 22-ohm resistor measured with the Boonton measured 60-pF inductive. With the homebrew bridge it measured 130-pF inductive. From the worked-out formulas in table 1 you see that resistor (as measured by the Boonton) results in an actual impedance of 21.82 ohms. As measured by the

like to correct the bridge to read out very closely with a laboratory bridge. In this way a low-impedance load could be measured and the nearly correct reactance could be found that would make the network a pure resistance. In fig. 8 resistors of different manufacturers, in 1/2-, 1- and 2-watt sizes and all values

from 18 to 2200 ohms were checked on the Boonton and on the homebrew bridge. In all cases lead length was held to one-half inch. The line marked *A* is the results with the Boonton. Incidentally, the slope of the lines implies a constant series inductance of .0268 Microhenry; this is the resistor lead inductance.

You can construct a similar correction graph for your bridge if you wish. On a piece of 2x2 cycle log graph paper lay out the curve *A* line from a point 16.5 ohms on the top line to a point at 165 ohms on the bottom line. Now, select a number of composition resistors from your junkbox — values from 22 to 100 ohms. Carefully measure these on your bridge at 14 MHz using half-inch leads protruding from points of contact with the load jack. The C_p values are the values of capacitance which are the *difference* between the *initial balance* value and that obtained in actually measuring the resistor. The R_p values are those read off the R dial when bridging the resistor. Plot these values for a number of resistors. Then draw a line parallel to the *A* line and through the approximate average point of your values.

To use the chart refer to fig. 8: That chart says that at 55 ohms the Boonton would read 9 pF, but mine would read 22.5 pF. If my R_p reading is 55 ohms, my C_p reading will probably be about 13.5 pF too high. Incidentally, as you tinker with this bridge, perhaps measuring the same resistor on each band, you may suspect something amiss when you consistently get the same value for C_p. You might suspect this number to be frequency dependent, but it is not. If you ran that same resistor through the range on a Boonton, you would find a slight downward shift in R_p as you go from 3.5 to 28 MHz (perhaps shifting from 28 to 27 ohms) but the C_p value would consistently hang in there.

A load which is not nearly pure R (as these are) will show the expected variations in C_p with frequency. If you feel that the half-inch lead length is gilding the lily, first bridge the resistor with full lead length, then repeat the measurement with half-inch leads. Don't be surprised if the C_p value of a 27-ohm resistor shifts as much as 50 pF. This effect is most noticeable at low resistances (50-ohm lines, etc.) and should serve as a warning of what can happen with sloppy clip lead lashups. I've seen otherwise competent engineers and technicians completely blow an rf measurement due to lack of feel for these little details.

If you are an engineering student, you will find this little bridge a valuable help in getting acquainted with rf measuring techniques. The Boonton RX meter goes all the way to 250 MHz, so you can appreciate the fact that it can get pretty hairy at times to get valid measurements.

using the bridge

Now, let's look at the business of making sense out of the numbers you get from your bridge. Table 1 is a set of formulas with worked examples to give you another method beside Smith charts as explained by W2CTK. The formulas assume F = MHz, R, X and Z in ohms, C in pF. you may wonder what ω represents. It is $2\pi F$. A handy little trick in amateur work is to remember that, for the low end of the bands, on 3.5 MHz, ω = 22, at 7 MHz, ω = 44, at 14 MHz, ω = 88, 21 MHz, ω = 132, 28 MHz, ω = 176.

The homebrew bridge reads a load in *parallel* values. The example in table 1 is for an almost pure resistance. It will be seen that the bridge read a 22-ohm resistance with a 56-pF capacitor in parallel. Actually, the 56 pF was in an *inductive* direction. This is the same as saying, "What coil in parallel with 56 pF will be resonant on 14 MHz?" You could find this on an ARRL Type A Lightning Calculator in less than 5 seconds. In the same length of time the calculator also says it would be a half-inch diameter coil, 1-3/4-inches long, 28 turns. You *know* a coil of those dimensions wouldn't likely be inside that resistor! If you are completely unfamiliar with the calculations you might be inclined to think it is a wire-wound resistor.

However, look at table 1 and find the

parallel reactance; it is 203 ohms. Such a large value of reactance in parallel with such a small resistance will hardly affect the resistance. It is the same as hanging an rf choke across the end of a pi network for safety. The pi net sees only the 50 ohms. If you continue down to the first calculation for impedance you see the carried out all your calculations to a number of significant figures both impedances would have come out identically. This last set of values sounds more reasonable! In fact, you actually do have a 21.74-ohm resistor in series with a pair of tiny coils (the resistor leads).

Now, suppose you had been using this

table 1. Mathematical formulas which may be used with the impedance bridge if you do not use a Smith chart. In all examples f = 14 MHz, Rp = 22 ohms, Cp = 56 pF and $\omega = 2\pi f$.

1. Parallel reactance $Xp = \frac{10^6}{\omega Cp} = \frac{10^6}{6.28 \times 14 \times 56} = \frac{1\,000\,000}{4928} = 203$ ohms

2. Parallel inductance $Lp = \frac{Xp}{\omega} = \frac{203}{88} = 2.30$ microhenries

   or $Lp = \frac{.159160\ Xp}{F} = \frac{.159160 \times 203}{14} = \frac{32.3}{14} = 2.30$ microhenries

3. $Q = \frac{Rp}{Xp} = \frac{22}{203} = 0.1083$

4. $Q^2 = 0.0117$

5. Absolute Impedance $|Z| = \frac{Rp}{\sqrt{1 + Q^2}} = \frac{22}{\sqrt{1.0117}} = \frac{22}{1.008} = 21.82$ ohms

6. Equivalent series resistance $Rs = \frac{Rp}{1 + Q^2} = \frac{22}{1.0117} = 21.74$ ohms

7. Equivalent series reactance $Xs = \frac{Xp\ Q^2}{1 + Q^2} = \frac{203 \times 0.0117}{1.0117} = 2.347$ ohms

8. Equivalent series inductance $Ls = \frac{Cp\ Rp^2}{(1 + Q^2)10^6} = \frac{56 \times 22^2}{1.0117 \times 1\,000\,000} = 0.0268$ microhenry

9. Absolute Impedance $|Z| = \sqrt{Rs^2 + Xs^2} = \sqrt{21.74^2 + 2.347^2} = \sqrt{478.1} = 21.85$ ohms

10. Equivalent series capacitance $Cs = Cp\ (1 + \frac{1}{Q}2) = 56\ (1 + \frac{1}{.0117})$

    $= 56\ (1 + 85.4) = 56 \times 86.4 = 4838.4$ pF

    or, $Cs = \frac{159160}{F\ Xs} = \frac{159160}{14 \times 2.347} = \frac{159160}{32.858} = 4840$ pF

total impedance of the 2.3 microhenry inductance in parallel with a 22-ohm resistance is 21.82 ohms. The coil hardly made any change.

Any parallel network can be converted to an equivalent series network or vice versa. From table 1 you see that an equivalent series resistance of 21.74 ohms in series with an equivalent series inductance of 0.0268 microhenries results in an impedance of 21.85 ohms. Had you bridge to measure a wide-band balun for your antenna. Suppose the balun was supposed to look like 22 ohms. Unless you have a feel for the effect of reactance you might have felt the balun was junk, while, in fact, it was almost a perfect match! For that reason it is worth taking the time to work out enough examples to get familiar with the arithmetic. Then, since the little bridge is quite accurate, you can be very sure of your results.

The impedance was figured in two ways to provide you with a means of back-checking your work. Whenever you convert from parallel to series values it will be worthwhile to use this as a proof that the series values were correctly figured.

How else can you use these numbers? Suppose you made a mobile antenna out of some odd scrap. You could end up with somewhat similar numbers to those

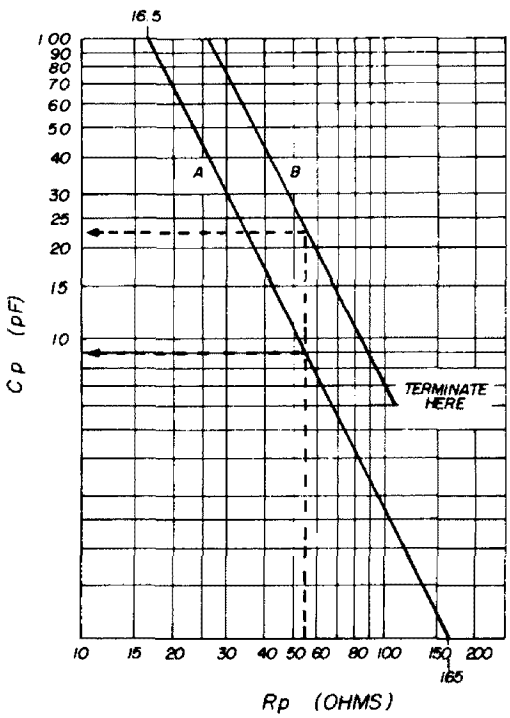


fig. 8. Calibration curves for the homemade impedance bridge (B), and a commercial Boonton 250 RX meter (A). These same curves can be plotted for your own bridge. See text.

in the example. If the antenna looked like 22 ohms in parallel with 56 pF *inductive,* you could make it 22-ohms resistive by merely hooking a 56-pF capacitor from whip to ground. If it looked 56-pF *capacitive* you could hook a 2.3-microhenry coil between whip and ground. Or, you could use the equivalent series values. If it showed 56-pF *inductive* on your bridge you could put about 5000-pF in series with the lead from the whip to the coax. If it looked *capacitive* you could put a 0.0268-microhenry choke in series from the whip to the coax. Of course, this example is pretty ludicrous, since it would be foolish to correct such a tiny inaccuracy. However, it gives you an idea of how very close you can work in antenna matching with this simple device.

applications

Another application would be to transform that 22-ohm whip to 50 ohms to match a transmission line. You might look up the arithmetic of L networks in the ARRL Handbook, but perhaps you are a bit unsure that the values you figured were correct. In such a case you can use the bridge backwards: Connect the L network to the whip, make the initial balance on the bridge, and connect it to the 50-ohm side of the L net. Set the R dial of the bridge at 50 ohms. *Don't touch the bridge!* Diddle the L network until the bridge nulls.

Perhaps you have an 80-40 dual dipole and want to include a balun to divorce your line from the antenna, but you can't afford the cost of the big toroid. Take a fat loopstick from a junked BC set and wind your 1:1 balun on it. Of course, you have no idea of the characteristics of the loopstick material. Connect one end of the balun to the bridge and solder a 51-ohm composition resistor to the other end. Vary the number and spacing of turns until it looks good. Remember that with a 51-ohm impedance it takes a fair amount of reactance to ruin the final impedance. Once the balun is done, put a dab of epoxy on the wire ends to hold them in place and you are in business at practically no cost.

Perhaps you built a little synchrodyne receiver and used an unmarked surplus toroid of unkown characteristics for an antenna coil. The set works fine except that when it's connected to a 50-ohm line it went dead. Don't spend a lot of time guessing the proper turns ratio! Perhaps the IC you were using called for 400 -ohms for best noise figure. Put a 390-ohm resistor across that winding, then bridge the antenna winding. When you get it around 50 ohms, non-reactive, put the receiver on the 50-ohm line and it will perform.

Perhaps you built a kilowatt linear from your junkbox. You carefully design-

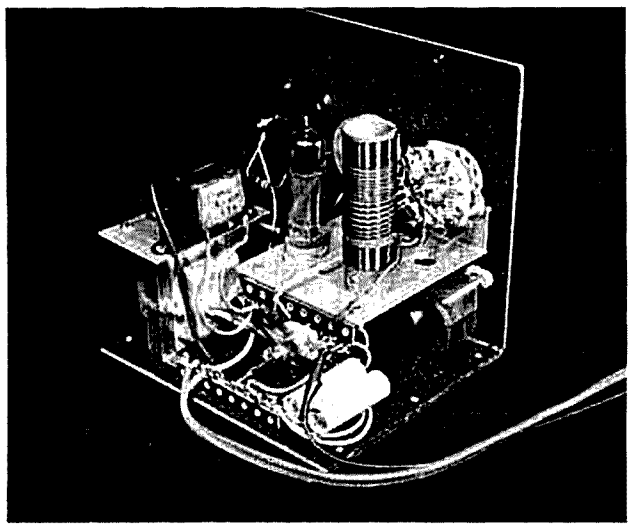
ed the pi network, but aren't sure the taps are correctly set on the coil to provide the proper plate load to the tube. Leave the linear turned off. Connect a composition resistor of the value you chose for the correct plate load from plate to ground. If the pi net is far from frequency you might not be able to bridge it, so set if for resonance with a grid dipper. Now look into the 50-ohm end of the pi net and adjust it to see if it will show 50 ohms with the capacitor values you calculated. If it does, you can be sure the plate sees the same load as the one you used from plate to ground. This will insure that the efficiency is up to the point you had calculated.

Of course, the impedance bridge is really useful in antenna work, but there are a few odd things you might run into. In the first place, if you bridge the coax at the transmitter end, it doesn't say that the antenna impedance is the same. Only if you take the time to bridge a good half-wave line (or multiple of half waves) and connect the end of that line to the antenna — will you get a valid reading.

W2CTK describes bridging a half-wave line. Essentially, make up one end of the coax, couple it to the bridge, terminate the far end with a 51-ohm resistor, check and snip, reterminate, check and snip until it reads 51 ohms, non-reactive point. Snip tiny pieces off so you don't whiz by!

If you bridge high-impedance antennas, you may run into certain quirks. My long wire showed the usual. The bridge was connected securely to the ground line coming in the basement window. This is a piece of number-4 wire about 3-feet long going to a ground rod outside the window. At that point it branches out with about 1000-feet of buried radials plus another hunk of number-4 to a second ground rod about 10-feet away. On 10 and 15 meters the bridge showed a lot of hand capacitance, and the null was quite dubious. The solution was to take the bridge outside the house, lay down a few strips of broiler foil on the ground, clamp them to the ground stake, clamp the bridge to the foil as close to the stake as possible, then lie down on the foil to put myself as near ground potential as possible. Now I had good nulls and accurate measurements. That 3-feet of number-4 lead was enough to put the bridge *above ground.*

Another odd factor showed up on 80 meters. The null would not go to zero and it wasn't hand capacitance. I shut off the bridge driver and the meter still read upscale. I soon realized that this antenna was quite nearly a quarter-wave long at the frequency of the local BC station. Measuring with my rf voltmeter showed a steady 200,000 microvolts of signal. *That* antenna makes a good one for checking cross modulation of receivers.



Construction of the rf power source.

conclusion

In conclusion, I feel that this instrument is an extremely useful device for the hamshack. With just a little care in construction and calibration it is remarkably accurate for its simplicity and low cost. I hope that the additional information presented here will encourage others to duplicate the device so that W2CTK can be properly credited with the importance of his little gem.

reference

1. Henry S. Keen, W2CTK, "A Simple Bridge for Antenna Measurements," *ham radio*, September, 1970, page 34.

ham radio

40-meter log-periodic antennas

How to design and build simple, high-performance wire beams that provide 8- to 10-dB gain and low swr over the entire 40-meter band

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In a previous article I described two fixed log-periodic antennas of the doublet type which cover the complete frequency range from 14 to 30 MHz.[1] These two antennas are suitable for use on 10, 15 and 20 meters. After that article appeared I received a large number of inquiries from amateurs who were interested in building a similar antenna for 40 meters. Although I didn't have a special requirement for a 40-meter beam, the possibility of building a log periodic for 40, 20 and 15 was interesting and presented a challenge.

Since there was space available between some of the high pines and cedars on my property, the large log-periodic beam shown in **fig. 1** was assembled. To reduce size and weight as much as possible, this log periodic is a *skip-band* type with a portion of the frequency range between 7.0 and 14 MHz deleted. Since this portion of the spectrum is not assigned to the amateur service, it is easily deleted and reduces the number of elements and the length of the antenna.

My available space only allowed a maximum antenna height of about forty feet, so I suspected that the beam would have little gain on 40 meters. However,

after running on-the-air tests for a few days, practically all the stations off the front of the antenna reported 8 to 10 dB gain from the 40-meter log periodic as compared with my non-gain 40-meter doublet. Reports from stations off the back of the beam indicated a front-to-back ratio of about 10 dB (signal strength

but still retain the gain indicated by the 40-20-15 log periodic.

The first 40-meter beam I built used six elements; the gain of this antenna was about 10 dB. Another log periodic was then built with only five elements; signal reports still indicated the same forward gain.

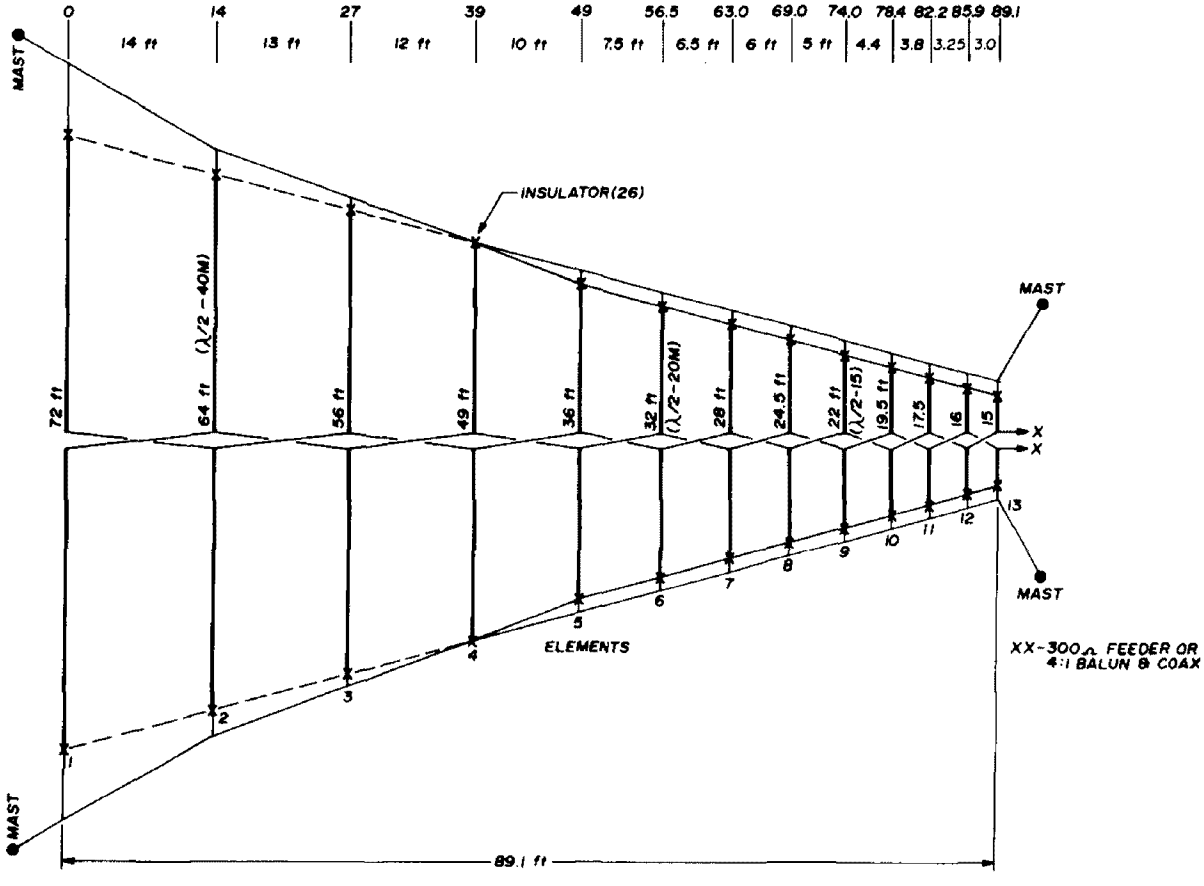


fig. 1. Three-band log-periodic antenna for 40, 20 and 15 meters. Forward gain of this antenna is 8 to 10 dB.

about the same as the non-gain antenna). Gain on 14 and 21 MHz is about 10 dB, and front-to-back ratios are about 15 dB, very similar to the three-band log-periodic previously described in *ham radio*.

forty-meter beams

Since many of the operators who work 40 meters are not interested in the higher bands, I decided to try a log-periodic for single-band use on the 7.0-MHz band. This would use a minimum number of elements and reduce the overall length,

These signal reports were very encouraging, but I wanted to obtain more accurate data, especially front-to-back ratio and side selectivity. As two high trees about 250-feet apart were available to the east of the three-band log periodic shown in fig. 1, I decided to put up a 5-element, single-band, 40-meter log periodic. Since only two high trees were available, I decided to use an inverted-vee configuration (center high, element ends low).

This arrangement would probably also

be better adapted to the needs of the average amateur because the antenna could be erected with two telescoping TV masts. With the two masts 60-feet apart, a single 40-meter log periodic can be installed in a space 60-feet wide by 75-feet long.

With inverted-vee construction a center A line between the two masts supports the open center feedline and the

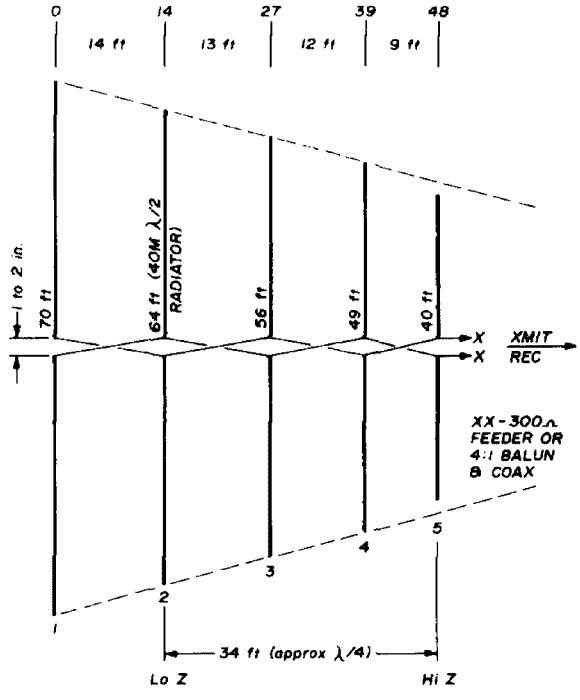


fig. 2. Simple single-band 5-element log periodic for forty meters. This antenna may by used in either the horizontal or inverted-vee configuration.

five elements. Each of the elements droop down about 45 degrees. The dimensions of this antenna are shown in fig. 2.

At my station the 40-meter inverted-vee log periodic was installed parallel to the 40-20-15 log periodic, at the same height, but about 150 feet away. By switching between the two beams I was able to determine what losses, if any, I had sustained by going to five elements in an inverted-vee arrangement.

I estimated that I would be lucky to obtain 5- or 6-dB gain with this shortened log periodic, but after about a week of on-the-air tests, I was convinced that there was little or no difference in forward gain between the two antennas. However, the front-to-back ratio of the five-element log-periodic was not quite as good as the 40-20-15 antenna. The signal reports that I received were usually confirmed by the S-meter at my station.

dual log-periodic beam

Another 5-element, inverted-vee, forty-meter section was mounted to the rear of the first beam and headed in the opposite direction (see fig. 3). The rear elements of the two back-to-back log periodics were spaced about 20-feet apart. On-the-air tests indicated that forward gain was still 8 to 10 dB. Reports from stations to the rear or off to the side of the antenna are about equal to a non-gain antenna, indicating a 10 dB front-to-back ratio.

construction

Monofilament fish line (40-lb test) is used for the element end insulators and also serve as the element support lines. These lines are brought down to posts or ground anchors. For one of the log periodics which I built I hung a line between several small trees and tied the monofilament end insulators to these. Square lucite insulators (fig. 4) are used as center insulators and support the open-wire center feeder.

The most recent log periodics which I have built have eliminated the center A

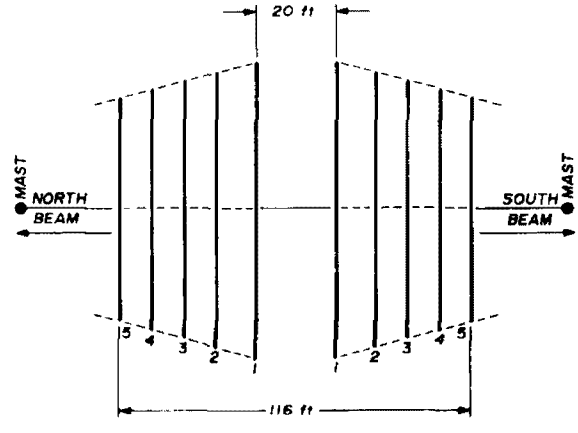


fig. 3. Single-band 40-meter log periodic for north-south communications consists of two back-to-back 5-element antennas (see fig. 2). This antenna, when used in the horizontal configuration, 50-feet above ground, provided 12-dB forward gain.

line. In these antennas the center is supported entirely by the center feeder. However, special precautions are required during assembly if you use this method of construction; the A line is easier if you are building your first log periodic.

It will be noted from fig. 2 that the open-wire center feedline between the second element and the rear element is approximately one-quarter wavelength long. The feedpoint impedance at the center of the front element is 100 to 300 ohms; the quarter-wave line acts as a transformer between the low-impedance at the center of the second element and the feedpoint at the rear. Three methods of feeding the antenna are suggested:

1. Open-wire tuned line with an antenna tuner in the shack.

2. A broadband 4:1 balun between the log-periodic feedpoint and a coaxial feedline to the shack.

3. 300-ohm TV twin-lead from the log periodic to the shack with a 4:1 balun at the shack with coax to the transmitter. I used this method of feed since my other log periodics are fed with 300-ohm tv line.

performance

Compared with my center-fed 40-meter doublet which is 50-feet high the inverted-vee log periodic consistently provides signal reports 8 to 10 dB better. Comparison with a commercial trap vertical mounted on the roof of the house indicate essentially the same thing. These are not just spotty tests, but are consistent, reliable reports from several old acquaintances in Florida with whom I have been working on 40 meters for several months. They were quite familiar with my normal signal on 40 using the two non-gain antennas before I put up the 40-meter log periodic.

After several weeks of testing I decided to increase the height of the 5-element south beam to at least 50 feet and use horizontal elements rather than the inverted-vee configuration. The results from this change were hard to believe. The signal reports were consistently at least 12 dB better as compared to the doublet, also 50-feet high. All signal comparisons were made when fading was at a minimum and there was no skip.

The front-to-back ratio of the horizontal log-periodic antenna is approximately 12 dB. This is near the possible maximum for log periodics built and tested here over the past two years,

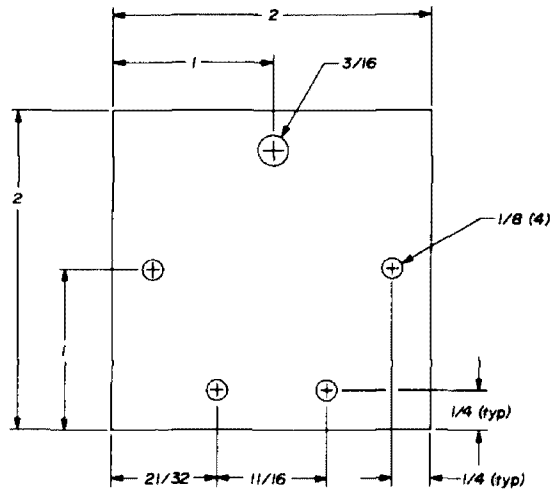


fig. 4. Center insulators for the log periodic antenna. Material is ¼" lucite or plexiglass.

reports from the back and sides are about the same as with the doublet I use as a standard.

As with any horizontal log periodic, the forward lobe is broader (about 100 degrees) than with most beams. This is good when you must use a fixed beam to cover a certain part of the country. Considering that most 40-meter beams are limited to 3- to 5-dB gain, the greater gain of the log periodic is certainly worth considering if you have the necessary space to put one up. By doubling the element lengths and spacing distances it should be possible to make a very excellent 80-meter beam.

I would appreciate hearing from anyone who builds and tests one of these antennas.

reference

1. G.E. Smith, W4AEO, "Three-Band High-Frequency Log-Periodic Antennas," *ham radio*, September, 1972, page 28.

ham radio

old 11-over-11 to the new box array. K2RIW in New Jersey recently gave me a report of S9+67 dB, so the antenna is really doing the job I wanted it to.

The antenna consists of four seven-element Yagis, spaced two half-waves. The four antennas are installed on the front of a square aluminum-angle frame which measures 27-inches per side. The antenna booms are 45-inches long. The entire array is fed with 75-ohm tv-type coaxial cable, which is lighter weight and has less loss than comparable 50-ohm coax.

construction

The driven element is made from 3/8-inch copper tubing. An 8-32 screw hole is drilled at the center of the element and a double-U clip is formed around the center of the element to keep it from deforming as you tighten up the screw. The U-shaped clips I used came from an old tv antenna. Wood dowels pushed into each end of the element and a solder-filled center would also keep the element from being deformed.

The phasing-line connections are made 1½-inch on each side of the center of the element (spaced 3 inches). A 6-32 screw hole is drilled all the way through the element and a 6-32 nut is soldered on one side. This can be done by polishing the copper with steel wool and sweat soldering the nut in place while it is being held temporarily with a 6-32 screw. Solder paste will speed things up; after soldering, the excess paste can be removed with lighter fluid. When connecting the phasing lines use two brass washers.

The reflector and directors were made from ¼-inch aluminum rod I salvaged from an old Telrex tv antenna. A hole is drilled in the center of each element to pass a 6-32 screw.

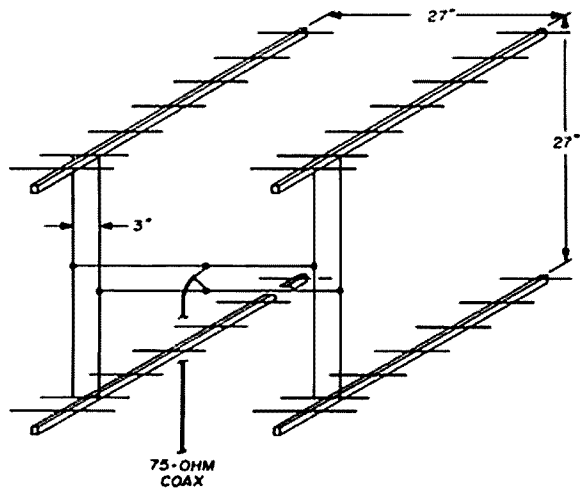


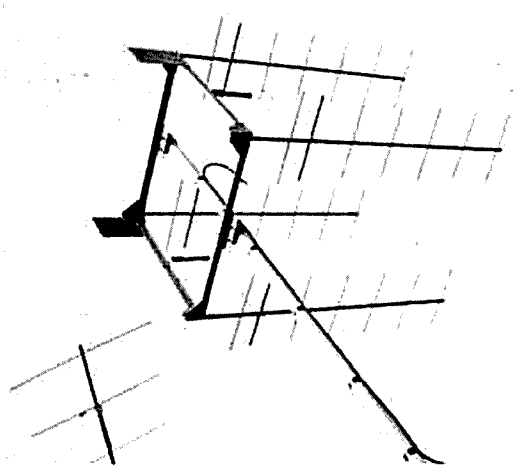
fig. 2. Phasing lines for the 28-element, 432-MHz quad array. Plexiglass or lucite spacers are used only at the junctions of the horizontal and vertical phasing lines.

The ½-inch square aluminum boom may require some searching. However, if you can't find one at your local building-supply store, it can be obtained on special order from firms that specialize in glass store fronts. A 6-32 screw hole is drilled in the boom for each of the elements. Then a slot is filed at right angles to the boom to hold each element in place.

the frame

The sides of the frame which holds the four Yagis are made from 1x1-inch aluminum angle, 1/8-inch thick. The material for the right-angle aluminum corner plates (see fig. 4) came from a local glass and store front company that builds six-inch aluminum columns.

When building the frame, attach the horizontal members to the corner angles first. Then the holes for the mast U-brackets are drilled. One of the flat faces of the aluminum angle faces toward



The 432-MHz quad-Yagi array installed at W3AED.

the Yagi antennas. The corner brackets are mounted on the outside. The individual Yagi booms are attached to the corner brackets with 8-32 screws, as shown in fig. 3.

The wind vanes on the back of the antenna shown in the photograph are

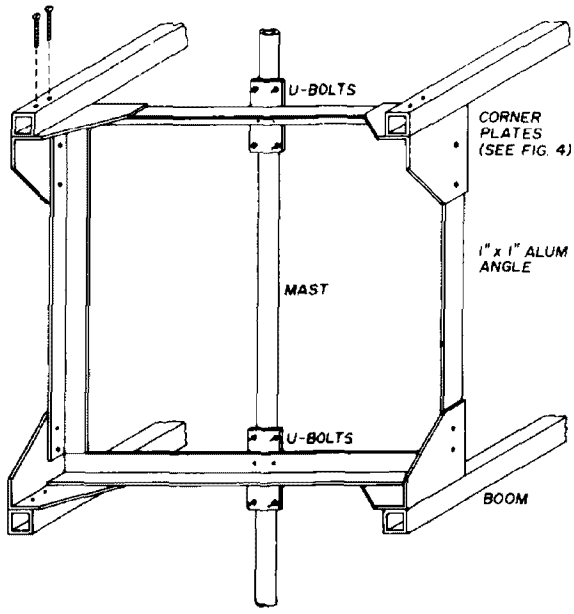


fig. 3. Construction of the rear-mounting frame. For 432 MHz the horizontal aluminum angles are 27-3/4" long; the vertical aluminum angles are 26½" high. For 1296 MHz the horizontal aluminum angles are 23½" long; the vertical aluminum angles are 22" long.

made from 5x8-inch pieces of aluminum sheet, although they may not be necessary if you don't live in a high-wind area. I also added a special plate to allow double U-clamps at each cross section of the box frame. Again, this is not essential, but I live in a hurricane area.

phasing lines

Each of the phasing lines is two half-waves long and made from number-14 wire, fig. 2. When using soft-drawn wire for this purpose I put one end in a bench vise and gave it a good pull. This gets rid of any future stretch and straightens the wire for immediate use.

Plexiglass spacers are used only at the junctions of the phasing lines. No additional support is required where the coaxial cable is connected. When installing the coaxial cable the shield braid should be cut back only far enough to spread the cable (1-inch maximum). After soldering, tape the connection well and tape the coax to the mast so the connection end of the coax faces down – this will help keep water out of the coax.

I used seventy-five feet of 75-ohm Jerrold CAC-82 coaxial cable with my array. This cable has a copper shield like RG-59/U and is packaged in 50-, 75- and 100-foot rolls.

1296-MHz array

Some readers might be interested in a four-bay, 13-element Yagi array I use on 1296 MHz which uses similar construction to the 432-MHz antenna. With this antenna and a three-stage preamplifier I have received signals from as far away as northern New Jersey and Norfolk, Virginia. Although I don't have the equipment to measure gain, performance appears to be within about 0.5 dB of the 1296-MHz Yagi described by W2CQH.[1] This indicates that gain is approximately 14 to 15 dB.

The boom for each of the 13-element

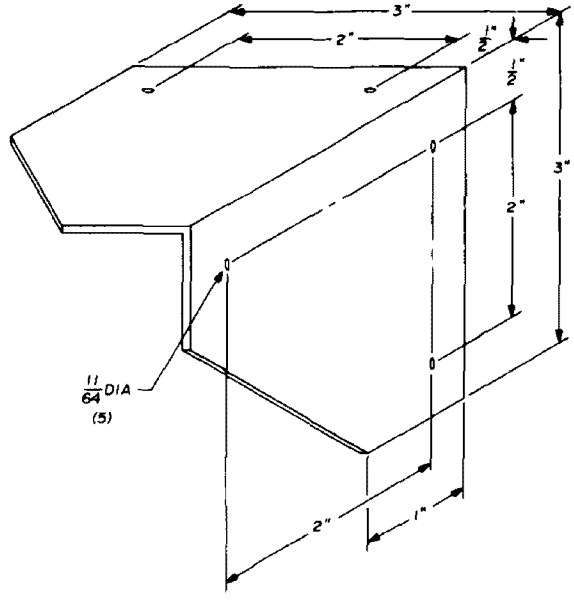


fig. 4. Corner angles for the rear-mounting frame are made from 3x3" aluminum angle, 1/8" thick. All holes are for 8-32 screws. These corner angles are installed on the outside of the aluminum angle frame.

Yagi beams is half-inch U-shaped aluminum angle, 0.050- to 0.075-inch thick. The driven element is brass. Other elements are aluminum rod. Element length and spacing is shown in **fig. 5.**

The upper and lower Yagis are placed face to face so that the phasing lines do not cross the booms. The conductors of the 300-ohm phasing lines are made from number-12 wire, spaced 1 inch. The vertical phasing lines are twisted at the center to meet the horizontal phasing line.

The design of the phasing lines on the 1296-MHz antenna is based on the fact that an impedance of 300 ohms, when connected to a 425-ohm matching section, exhibits 600-ohms impedance. When the two 600-ohm sections are connected in parallel, the input impedance is again 300 ohms. A 4:1 balun at the center feedpoint converts the 300-ohm input impedance to 75 ohms, a perfect match to 75-ohm tv type coaxial line. I used a 50-foot section of Jerrold 82 tv transmission line, which has a copper braid.

Although this 75-ohm tv coax comes pre-packaged in lengths of 50, 75 and 100 feet, any length over 50 feet is much too lossy on 1296 MHz. If you need more than 50-feet of transmission line at your station you should consider placing the first stage of your rf amplifier right at the antenna.

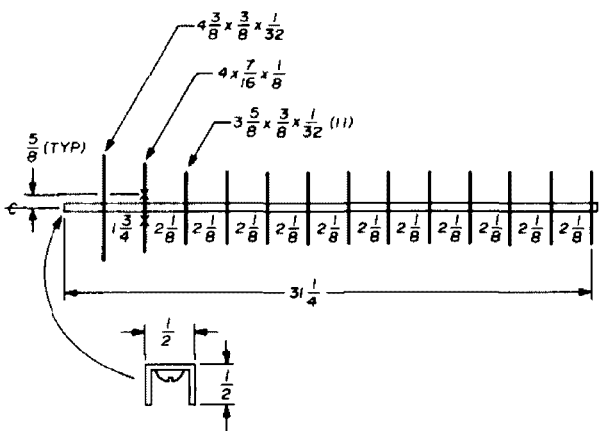


fig. 5. Construction of the 13-element, 1296-MHz Yagi beams used in the four antenna array. The individual Yagis are mounted on a square frame similar to that shown in fig. 3.

rf filter

A number of amateurs have had trouble on 1296 MHz because they did not place a 1296-MHz filter ahead of their rf amplifier. Many of the popular 1296-MHz preamplifiers run wide open and pick up a lot of undesired signals. Occasionally, they also oscillate when connected to a resonant antenna. The simple half-wave rf filter shown in **fig. 6** solves these problems.

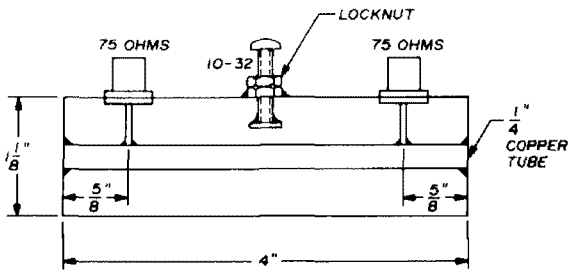


fig. 6. Half-wave 1296-MHz rf filter prevents problems with spurious signals and self-oscillation. For 50-ohm input impedance, connectors should be spaced ½" from the ends of the line.

The half-wave filter is built into a 4-inch long brass or copper box, 1-1/8-inches square. For use with 75-ohm lines the connectors should be spaced 5/8-inch from the end of the line. For 50-ohm lines this spacing should be ½-inch. By spacing one connector ½-inch and the other 5/8-inch, the filter can be used to match 50-ohm cable to a 75-ohm transmission line.

When the filter is inserted in the line, the 10-32 tuning screw is adjusted for maximum signal. The screw is then locked in place with the locknut. I am presently using this filter ahead of three rf stages with good gain and no oscillation.

I would like to take this opportunity to thank Cy Dirickson, W3BSV, who helped with the design and testing of these antennas.

reference

1. R. Fisher, W2CQH, "A Successful 1296-MHz Yagi," *ham radio*, May, 1972, page 24.

ham radio

antenna and feedline facts and fallacies

New experiments provide some interesting facts about old antenna and feedline fallacies

Carl C. Drumeller, W5JJ, 5824 NW 58 Street, Warr Acres, Oklahoma 73122

How many times have you heard discussions, often heated, about the following subjects? Does a dipole antenna fed with a coaxial cable radiate more from the leg fed by the cable's center conductor than from the one fed by the shield? If so, does the use of a balun at the feedpoint correct the imbalance of radiation?

Does a coaxial cable feeding a load mismatched to cause a 4:1 vswr have radiation from or standing waves upon its outer conductor?

Does a coaxial cable feeding a dipole antenna, with the feeder running at right angles from the antenna, have current on its outer conductor? If so, does the use of a balun at the feedpoint correct the condition of undesired radiation?

Does one leg of the dipole antenna plus a quarter-wave of the outer conductor of the coaxial feeder constitute a vee antenna, thereby causing a high-voltage rf potential to appear at the quarter-wave spot and at even multiples of it on the transmission line?

Well, I've heard these questions countless times and have even taken part in some of the arguments. The distinguishing factor about all these discussions is that no one really knew what the answers were. Oh, everyone was quite sure in his own mind that he knew the answers perfectly. But, and this is the point I want to make here, not a single person had ever taken part in an experimental determination of the truth or fallacy of the several contentions! Let's have an end to such a deplorable situation.

Three amateurs, W5TMY, W5KE and myself, each retired and blessed with ample free time, each with over thirty years of professional experience in the field of electricity and electronics, decided to do something about establishing demonstrable and reproducible facts relating to the disputed questions. My place was selected for the experimental setup as I have a backyard of ample dimensions, and my 50+ years collection of equipment and instruments contained everything we needed.

The basic equipment consisted of a Drake TR-4 transmitter, A Drake MN-4 matching device (helps minimize harmonics), 52-ohm coaxial line, a 7-MHz dipole antenna in the clear and low enough to permit an indicating device to be pulled along it, and a sensitive device for measuring relative amplitude of radio-frequency current (current, *not* voltage). See fig. 1 for details of this indicator.

The vinyl jacket of a portion of the coaxial cable was removed to permit

exploring the radio-frequency current on it over a half-wave expanse. The cable was run at right angles to the antenna and was at the same height as the antenna. Every effort was made to preserve absolute symmetry. A W2AU balun, especially made with a male coax connector at one end and a female connector at the other end, was used for altering the feedpoint circuit.

experiment 1

Purpose: To ascertain whether the currents in the two halves of a dipole antenna are equal when fed from an unbalanced source, a coaxial cable.

Procedure: The antenna was energized, the radio-frequency current indicator was adjusted for 45/50th full-scale deflection at the spot of maximum current, and the indicator was pulled by means of an attached string along the length of the dipole half. Without adjusting the indicator, the action was repeated on the other half of the dipole.

Findings: The maximum current in the half fed by the center conductor of the coaxial cable was approximately 40% greater than that in the half fed by the cable's shield.

Conclusion: Feeding a dipole from an unbalanced feeder results in an imbalance of currents in the two halves of the dipole, *probably* resulting in a distorted radiation pattern.

experiment 2

Purpose: To ascertain whether the currents in the two halves of a dipole antenna are equal when fed from a balanced source: a balun coupling a coaxial cable to the antenna's feedpoint.

Procedure: A balun was inserted between the cable and the antenna; then experiment 1 procedure was repeated.

Findings: The maximum current in each half of the dipole was very nearly equal.

Conclusion: The use of a balun helps to obtain balanced radio-frequency currents in the two halves of a dipole.

experiment 3

Purpose: To ascertain whether having a coaxial cable terminated in a mismatched load causes radio-frequency currents or standing waves to be present on the outer conductor of the cable.

Procedure: A length of 52-ohm coaxial cable, more than a half-wave long and having insulation stripped from the outer conductor, was terminated with four

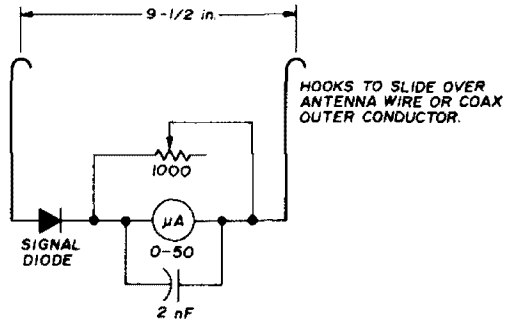


fig. 1. Construction of the sensitive rf current indicator.

52-ohm Termaline dummy loads, thereby causing a 4:1 vswr to be present on the line. The line was matched to the transmitter through an MN-4 matching device, and 200 watts of rf power was fed into the line. The sensitive rf current indicator was pulled along the length of the line.

Findings: No radio-frequency current was detected on the outer conductor on the coaxial cable.

Conclusion: Having a mismatched load does *not* result in the radiation of rf power from a coaxial transmission line. However, this does not imply that having an *unbalanced* load may not cause current to be present on the outer conductor.

experiment 4

Purpose: To ascertain whether there is rf current on the outer conductor of a coaxial cable directly feeding a dipole antenna.

Procedure: A length of coaxial cable, more than a half-wave long, stripped of its outer insulation, was strung at right

angles to the dipole and at the same distance above ground. The transmitter was loaded to 200 watts, and the sensitive rf current indicator was pulled along the transmission line. An indication of full-scale (but not pegging the meter) deflection was observed. In an effort to obtain a comparison of relative power at this maximum current point and that at the maximum current point on the dipole, the transmitter power output was reduced to a value too low to be read on the MN-4's rf wattmeter (probably less than one watt). Even at that low power it was necessary to severely shunt the microammeter on the indicator in order to prevent its being violently pegged by the antenna's rf. (*Note:* The point of maximum rf *current* was approximately one-half wave back from the feedpoint.) **Conclusion:** There indeed is rf current on the outer conductor of a coaxial cable having an unbalanced feed to a dipole. This current, however, is *very* small. The radiated power ratio, cable versus antenna, can only be estimated; probably it lies between 1:1000 and 1:10,000. The assumption that there is a high-voltage point on a cable's outer conductor a quarter-wave back from the feedpoint is *not* valid.

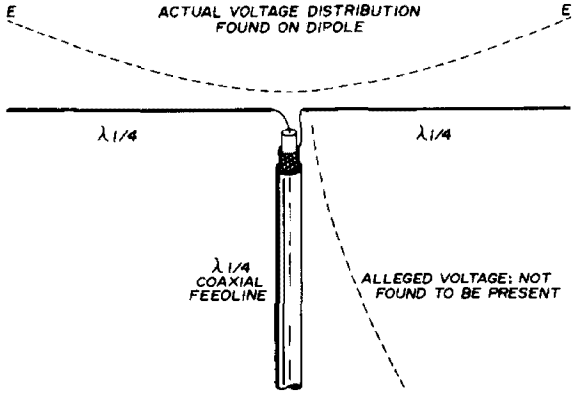


fig. 2. Rf voltage curves on the dipole and coaxial feedline were plotted as reciprocals of the current measurements.

experiment 5

This started out as a duplicate of **experiment 4** but with the balun in place. No rf current was detected on the cable's outer conductor; so the experiment was terminated.

the results

Let's evaluate the results. The use of a balun gives some small advantage, both in balancing the radiation from the dipole and in diminishing the radiation from the coax.

The assumption that a mismatched (mismatched, not unbalanced) load, which engenders a moderately-high vswr, causes standing waves upon and radiation from a coaxial transmission line is without basis. Somebody will have to come up with a new excuse for all those mysterious cases of TVI, "hot" microphones, etc., that have been explained off as, "standing waves on the transmission line."

A transmission line does not transform itself into a vee antenna consisting of a half dipole plus a quarter-wave of transmission line outer conductor. Again, some new excuses are needed.

For your own evaluation of these findings, please consider that the experiments were conducted under rigid control. All findings were duplicated by another person before being accepted. Two of the persons involved are retired electronic engineers, the other a retired electrical foreman; all are experienced radio amateurs. The instruments used, with the exception of the homebuilt rf current indicator, were high-grade, equal to what might be found in a laboratory. Every effort was made to have all findings factual – not to "prove" some previously-held assumption.

Although this article is being written by one person, I want to emphasize that J. D. Odneal, W5TMY, and Ellard W. Foster, W5KE, were equal participants in the project. It could not have been conducted without their co-equal participation.

If you have reason to doubt our findings, I beg you not to dispute them solely from a theoretical basis. Duplicate the experiment, then express your opinions!

ham radio

80-meter antenna for a small lot

If you live in a stucco house on a small city lot, here's how to install an 80-meter antenna

Alfred Stahler, W6AGX, 5521 Big Oak Drive, San Jose, California 95129

Housing tracts with small, uniform, close-spaced lots exist throughout the United States. In California the typical house on such a lot uses stucco construction. This is fine for the climate, but the galvanized-steel wire mesh used to support the stucco plaster is usually well grounded at each electrical outlet to the ac ground as well as to the cold-water pipe. Since this conductive, well grounded, two story, "shield can" rests approximately in the middle of the lot, it represents a non-ideal support for an amateur antenna; particularly a long, low-frequency antenna.

Faced with these facts, I tried to build an 80-meter antenna that would fit on the 60- by 100-foot lot, require no towers, be unobtrusive and be broadband enough for reasonable swr over the CW as well as phone portions of the band. I tried several configurations from U, L and V, ending up with the final Z shape shown in fig. 1. The desired broadband characteristics, in all cases, were dependent upon the "double bazooka" concept which combines the characteristics of a resonant coaxial length with a radiating dipole.[1]

Because of the odd shape due to the lot size restrictions and the influence of the stucco-house "shield can" I found little correlation between published theory and my experimental measurements. I proceeded on a cut-try-measure-cut-try-etc. basis until the trends yielded a reasonable antenna.

The physical layout of the antenna, house and lot is shown in fig. 1. The coaxial center section of the antenna runs along the peak of the roof and then down the roof lines toward the two supporting

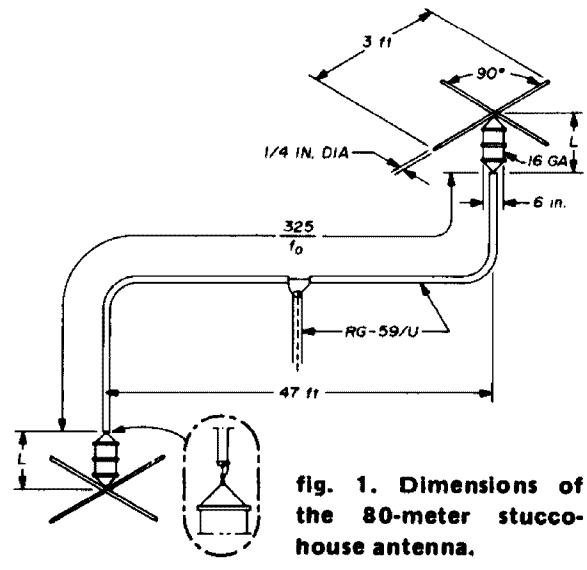


fig. 1. Dimensions of the 80-meter stucco-house antenna.

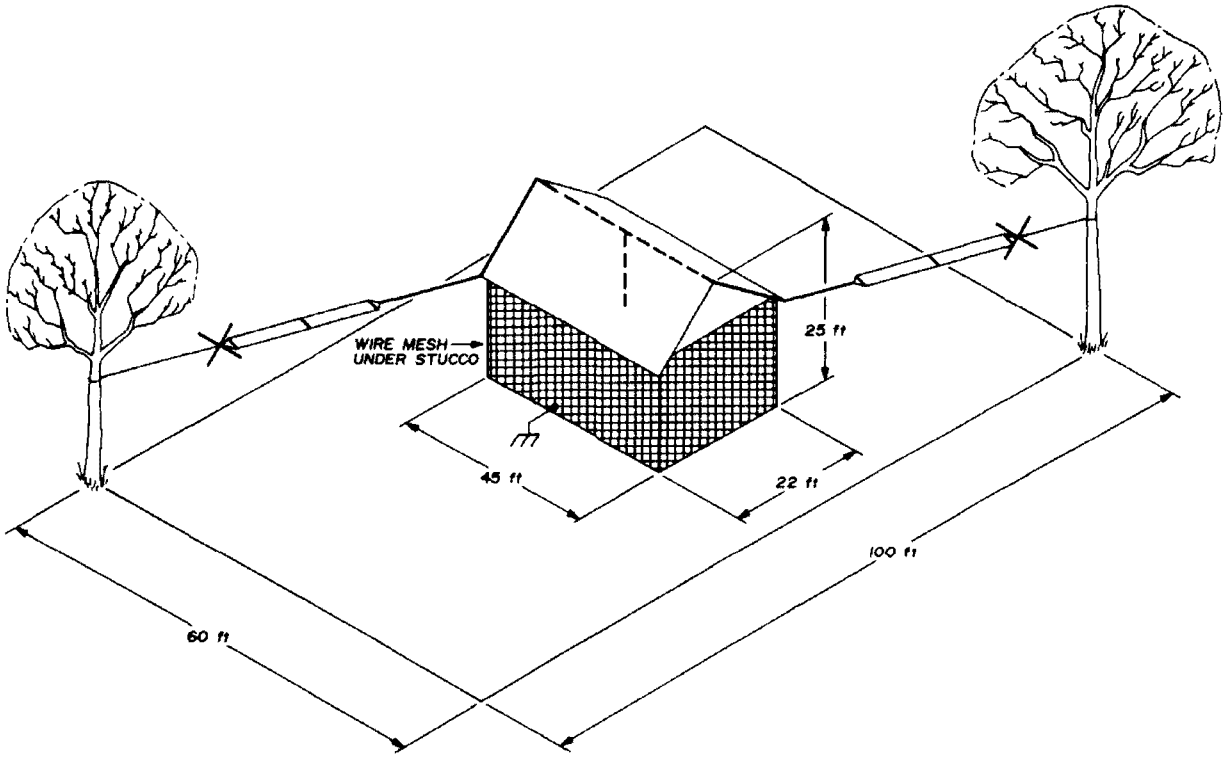


fig. 2. Installation of the 80-meter antenna.

trees. The coax feedline runs down vertically through the house to the transmitter. At the ends of the coaxial section are sections of open-wire transmission line with capacitance hats. The hats are made of two ¼-inch diameter, three-feet long, crossed aluminum rods. The dimensions of the antenna are shown in fig. 2.

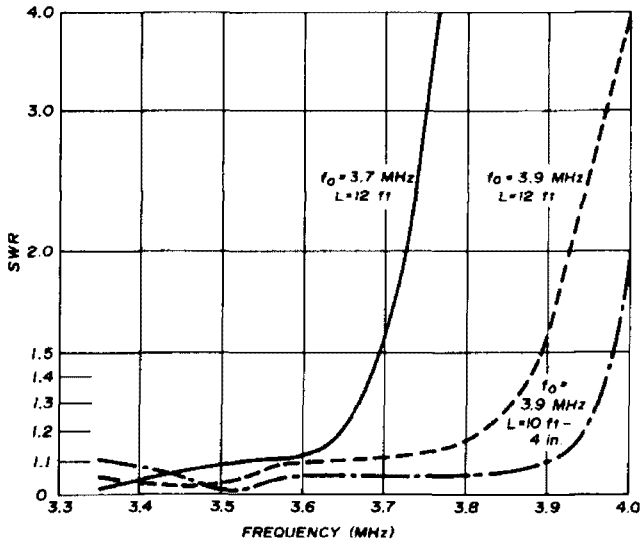


fig. 3. Standing-wave ratio vs frequency for different antenna lengths. The length of the open-wire sections at each end of the antenna is given by L; f_o is the design frequency in MHz where the length of the center section (in feet) is given by $325/f_o$.

Fig. 3 shows the measured swr vs frequency for several versions of this antenna. It can be seen that the influence of the Z-shape and the stucco house have drastically changed performance from that predicted in reference 1 for an isolated dipole. However, by trial and error the final geometry performs quite well over a large portion of the 80-meter band. Signal reports have been good, and the antenna seems to exhibit dipole-type directivity if the transmitting axis is assumed to be in a line with the antenna end points.

The rf voltage at the tips of the capacitance hats is very high so they should be placed well away from leaves, branches and people.

I hope that other amateurs who are restricted to life on similar house lots will be able to use this antenna configuration on 80 meters. Perhaps a similar arrangement could be developed for 160 meters.

reference

1. D. DeMaw, W1CER, *The Radio Amateur's Handbook,* 48th edition, A.R.R.L., Newington, Connecticut, 1971, page 368.

ham radio

simple antennas for two-meter fm

Here is a selection of simple vhf antennas which can get you on the air at minimum cost

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The two-meter fm band is booming and the results are exciting, chiefly because of repeaters which make even crummy antennas acceptable for receiving and transmitting. Of course, if you have the money, you can spread yourself and buy commercially-made antennas which represent topnotch engineering design.

Looking through the Lafayette Catalog, for example, you can find a nice Ringo antenna which covers 135 to 175 MHz for $12.50. This is a vertical which may be used for mobile operation with the metal body of the car serving as a ground plane. It has a gain of 3.75 dB over an isotropic antenna. If there is one thing I've learned about verticals it is that a good ground system is necessary. However, if the thing is way up on a mast, far from ground, four radials will do the job. The radials should be slightly longer than the vertical radiating element, perhaps by 10%.

Since most mobile installations use vertical antennas it helps if the receiving antenna at the base station is also vertical. In some cases it doesn't seem to matter,

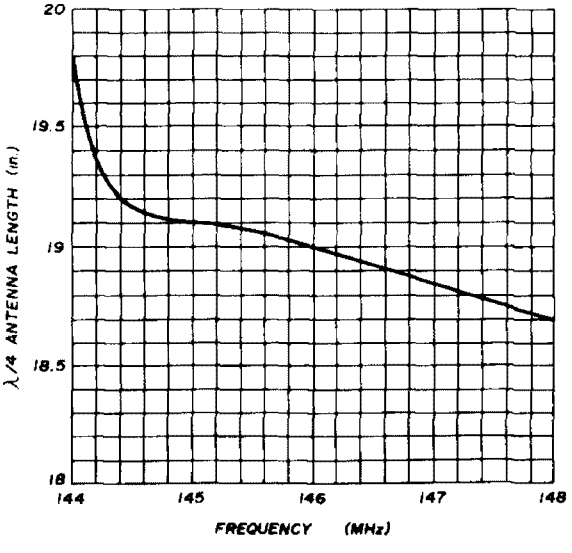


fig. 1. Length of quarter-wave two-meter antennas.

especially in hilly country or wherever there are numerous reflections which may change the relative polarization. The usual rule, however, is a vertical receiving antenna if the source transmitter uses a vertical.

In the *ARRL Radio Amateur's Handbook* halos and other weird horizontally-polarized antennas are shown, but hams don't seem to want to mess up their expensive autos with screwball antennas. The vertical is relatively unobtrusive, especially on two meters. I have seen some halos on masts mounted to bumpers on Volkswagens, and one guy even had a 10-meter whip on the front bumper of a VW.

The two-meter antenna can be relatively short. In a motel one ham took the oven grill out of the oven and used it as a ground plane for a small vertical with a magnetic mount on two meters, driving it with a Drake TR-22. He had a barrel of fun with it, using only a watt of rf power to get out. The repeater is a great thing.

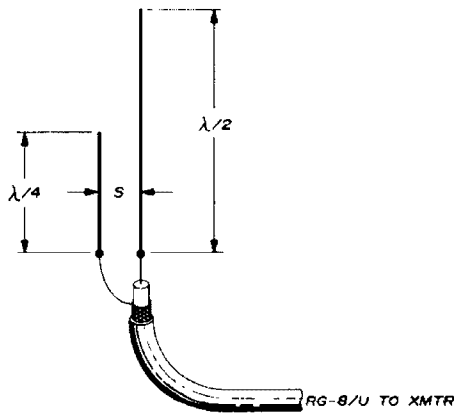


fig. 2. The vertical J-antenna is an efficient radiator. Spacing S should be 2 inches or less.

On page 429 of the Lafayette Catalog, Antenna Specialists has a roof-mount antenna for $22.95, net. It is equipped with a stainless-steel whip with copper and nickel plating for low rf resistance. For home use a 7-element Cushcraft beam goes for $13.95 net and gives a forward gain of 11 dB. That's a lot of gain, and it really makes a difference. An 11-element, 2-meter Yagi antenna, Cushcraft, has a forward gain of 13.2 dB and goes for $17.95. At these prices, it hardly pays to build one yourself. Yet, the ARRL Handbook is full of material that assumes you are a radio engineer, master mechanic and have the facilities of a well equipped radio laboratory and machine shop. Some hams may have such facilities on their jobs, but access to them is a different matter.

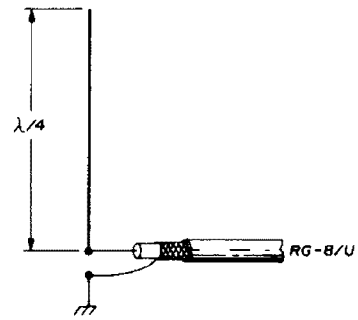


fig. 3. Simple quarter-wave vertical.

A 22-element, 2-meter Yagi, Cushcraft, has a fantastic gain of 16 dB and sells for $49.50 net. Of course, the thing has to be put up on a mast somewhere, which may be the chimney of your little house, or you can buy an expensive mast for the backyard. Being poor, I can only dream of such goodies.

Heavy-duty antenna rotors may run $69.95 or more, depending on what you select. A directional antenna must be rotated unless you are interested in operating in only one direction.

simple antennas

The simple quarter-wave antenna shown in **fig. 1** is from 18 to about 20-inches long for the 2-meter band. Around the center of the band, 146 MHz, which is where most of the activity seems to be, a quarter-wave is about 19 inches. You can use brass rod or even copper

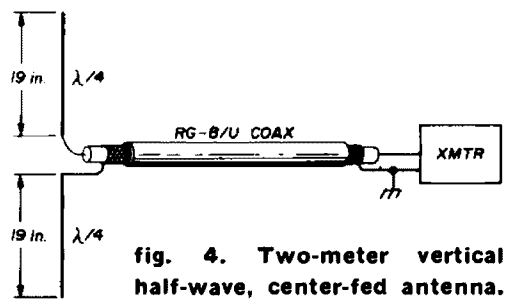


fig. 4. Two-meter vertical half-wave, center-fed antenna.

tubing, and no. 14 hard-drawn enamel wire clipped to a 2x4 could even be used. Such antennas, while comparatively crude, are inexpensive and they work.

Other simple vertical antennas are shown in figs. 2 and 3. These require good grounds if not far above earth, not more than 10 feet, and ground-plane radials if they are mounted on a mast. By going to half-wave center-fed antennas the ground troubles are reduced, although any antenna generally works better over low resistance ground.

A simple beam is the Lazy-H (see fig. 6). This antenna works well on the high frequencies as well as on 6 and 2 meters. It can be put up in a horizontal plane above the house on a boom and rotated.

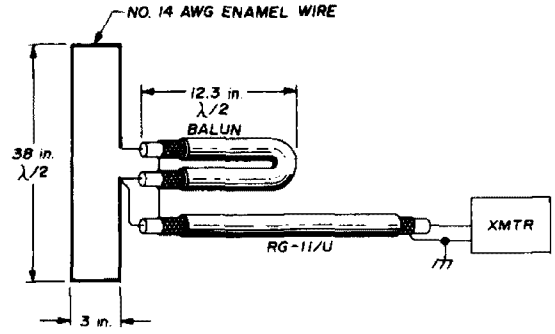


fig. 5. Vertical half-wave folded dipole for two meters.

The Lazy-H is one of my favorites. It can also be mounted in a vertical plane on the wall of a room in an apartment. Ordinary zip cord has been used for both the elements and the feed. However, fairly fine no. 22 enamel wire can be used for the elements and the feedline can be comparatively light-weight RG-59/U. The impedance at the feedpoints should permit using 50- or 75-ohm coax cable directly. If necessary, a quarter-wave matching stub may be used.

circular antenna

By bending some copper or brass rod into a circle you arrive at the antenna shown in fig. 7. This antenna can be used in a horizontal or vertical plane and fed with coax. The coax should be kept as short as possible to avoid losses which

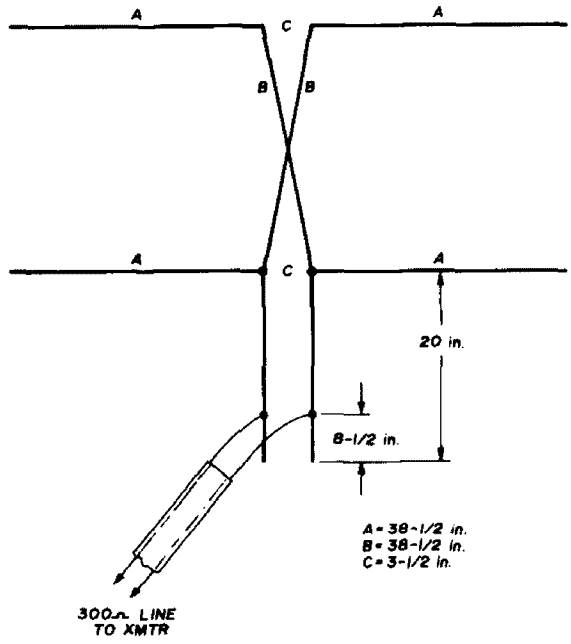


fig. 6. Directional Lazy-H beam for two meters may be either vertically or horizontally polarized.

tend to mount up at vhf. Some cable characteristics are shown in table 1. If you have an antenna with 16 dB gain and you lose 6 dB in the cable run you still have 10 dB left, but with a cheap vertical a 6-dB loss is intolerable.

rabbit ear antenna

The rabbit ear antenna, fig. 8, may be used as a horizontally polarized half-wave

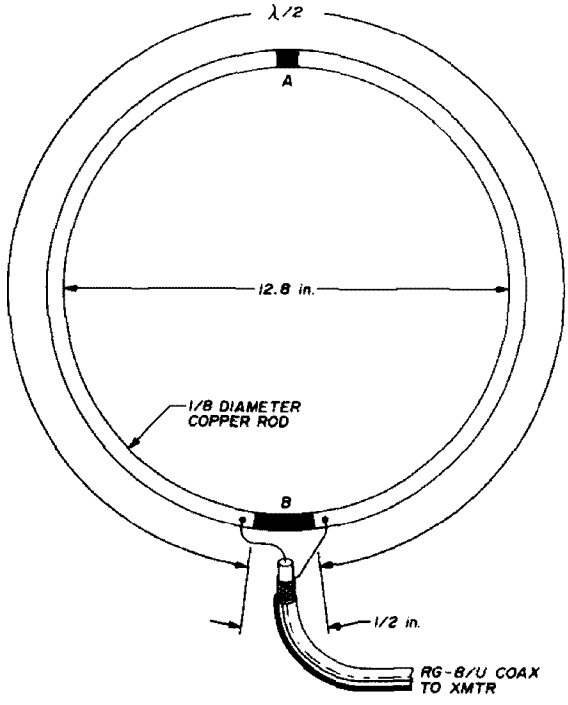


fig. 7. Two-meter circular rod antenna.

table 1. Vhf characteristics of common coaxial cables.

	impedance (ohms)	loss (dB/100 feet at 150 MHz)
RG-58/U	52	6.0
RG-59/U	73	4.5
RG-11/U	75	2.5
RG-8/U	52	1.8

antenna (19 inches for each element at 146 MHz). See fig. 1 for other dimensions in the two-meter fm band. Some of these tv antennas are well made and others are junky. Try to get a substantial, well-built one. The 300-ohm line may be clipped near the base, leaving tabs long enough to solder to a piece of 50- or 75-ohm coax.

The length of each element can be

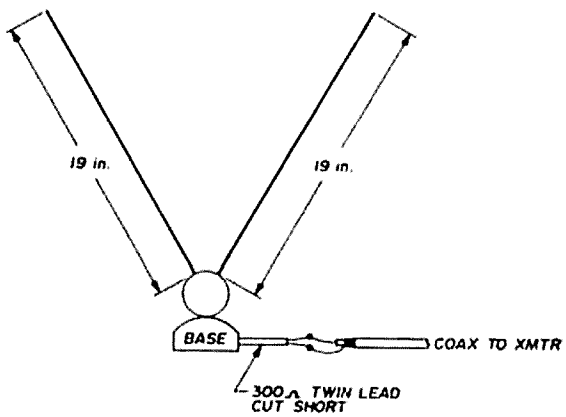


fig. 8. Television rabbit-ear antenna may even be used on two meters. The length of the elements is adjusted for resonance on the desired operating frequency.

adjusted with a tape measure or an engineer's scale. Each element can be tuned and resonated by the sliding action or trombone-like construction. This makes a simple and convenient low-power antenna. A Drake TR-22 will work into it, but don't try to feed a kilowatt to it unless you like smoke.

So, as the old saying goes, you pay your money and take your choice. If you are loaded, buy a commercial antenna. If you are watching the budget try one of the simpler types. They seem to work fairly well although some of them are not much for appearance.

ham radio

Proved in Commercial 2-way Now Available To Amateurs!



Larsen Mobile Gain Antenna 144-148 MHz

The result of over 25 years of two-way radio experience. Gives you . . .

- 3 db + gain over 1/4 wave whip
- 6 db + gain for complete system communications
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how to tailor your antenna for optimum performance

A few ideas for building simple antennas, and tailoring commercial ones to your individual operating habits

Albert F. Lee, KH6HDM, 204 Makee Road, Honolulu, Hawaii 96815

Tailoring your antenna for your particular area and getting it to work properly can do wonders for your enjoyment of amateur radio. If "clothes make the man," and a tramp with a shave, a warm bath and a new suit looks like a company first vice president, then you can do a lot better by tailoring your antenna for that proud feeling.

We, nearly a half-million hams, live in a myriad of locations and conditions. Few of us have found the antenna which suits ourselves let alone the rest of the fine people in this hobby.

Particularly difficult is the challenge to the apartment dweller. There are an increasing number of tract and land-rental legal restrictions written into deeds and rental agreements. More than a few communities have very restrictive laws on antennas. To all of this add the watchful eye of your spouse and your neighbors' objection and you will see just a few of the problems you face when choosing an antenna.

Space is the prime guide when choosing an antenna, with cost and appearance being secondary factors.

the dipole

This simple antenna consumes little space, is the least expensive and can be made almost invisible. By fanning out the individual wires, you can obtain satis-

factory multiband operation. The swr is most simple to bring to unity by adjusting the leg lengths and the angle between the elements. Performance can be enhanced by the addition of a balun.

Placing the dipole in a position free of wave obstruction and at least one-quarter wavelength above ground can give some real fine performance. The preferred height above ground is one-half wavelength, but we don't always achieve perfection in our antenna tailoring.

beams

Utilization of the driven element, reflector and director is best seen in the beam antenna where you use aluminum tubing, coils and a rotating mechanism. You can handily operate a beam on one band or a combination of three. The most popular arrangement is the 10-, 15- and 20-meter combination. The two great advantages of such a beam are the gain and the directivity of the signal, which cuts out considerable interference. A Yagi beam, resonated to the correct frequency, balun and half-wave transmission line is the best antenna system which I have tried. This is the experience of very many other hams, too. A good, well-tailored Yagi is hard to beat.

Many amateurs are short on aluminum tubing and fittings, coils, rotators and the tailoring ability to make such an array as described above. Fortunately, there are two types of wire beams which are easy to build and inexpensive as well. The first is the three-element 20-meter wire beam shown in fig. 1. It consists of a driven element, with balun, placed one-half wavelength above a reflector of wire, which is 5% longer than the driven element. This will radiate a greater signal at right angles to the driven element. This may be increased by placing a director one-quarter wavelength in your preferred direction. The director should be 5% shorter than the driven element.

With a grid-dip meter and an antenna bridge you may adjust the driven-element to director spacing to the desired operating frequency and input impedance, with the gain checked with a ham friend a couple of hundred miles away. This antenna will really whistle.

The second wire beam is a two-element job with a little different and easier tailoring as shown in fig. 2. The driven element is made of light-weight aluminum tubing. I obtained 1/2 inch

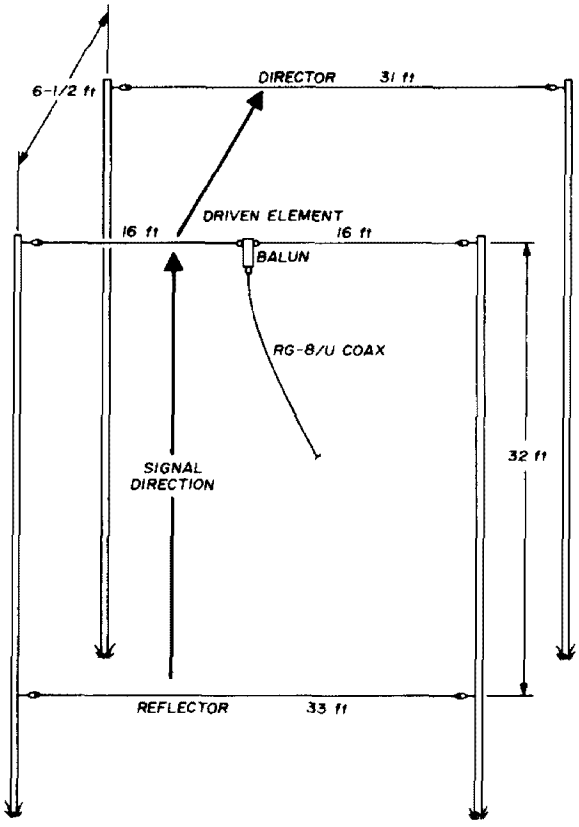


fig. 1. Three-element 20-meter wire beam with stacking of driven element on reflector. Director is parallel to the driven element.

surplus aluminum tubing which was made so that one six-foot length would slide into another. With these sections I made two such lengths, each 16-feet long. These two elements are separated with a piece of wooden dowling with the stabilizing bolts attached to each side of the balun. The RG-8/U coaxial transmission line was cut to one-wavelength long at the operating frequency of 14.270 MHz and attached to the transmitter.

This driven-element dipole arrangement is attached to a bamboo spreader supported by a nylon bridle. The director

wire is 31-feet, 4-inches of number 14 copper wire, placed 6½-feet toward my desired direction. I raised this array eight feet above the roof and my signal doubled.

Now before I could brag about this, my happiness was dampened a bit when I heard the loudest station on 20 meters telling his friend in New York about his 20-meter wire beam made with three elements — all of number-25 wire. I was

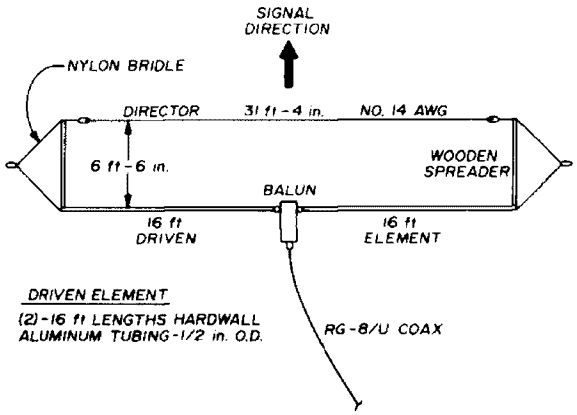


fig. 2. Wire and aluminum tubing beam for 20 meters.

reading all this on the back of this beam so I know, now, he is a better tailor than I.

verticals

The vertical antenna is dandy for limited space as long as you observe a few good tailoring rules. Verticals require a good clear space and perform best with a good ground plane or radial system. The better these conditions are met, the better the function. However, it is a surprise how well a vertical will work with less than ideal conditions. A vertical cut to the desired frequency with one wire to a water pipe, ground stake or even a base of random chicken wire, will occasionally do a passable general and DX job. This is almost as good as a system with a system of radials or a ground-plane.

Single-band verticals can be phased in the same manner as beams for gain and directivity. I find that the vertical is the easiest of all antennas to resonate to frequency and impedance match. Furthermore, these adjustments can be made without climbing towers or raising and lowering the elements. Minimal work gives maximal success.

quads

This closed loop and its cousin, the Delta Loop, have a full-wave driven element and reflector working to give a fine gain and directivity. Performance is great with a quad but I can personally vouch for the large size and fragility of this type of antenna. Here is a place for you to tailor yourself a real fine antenna if you have the space and skills which I found I lacked.

rhombics and long wires

The rhombic is beyond the usual ham as it takes a lot of space which most of us don't have. I saw a grand rhombic in use on field day — my only observation was that they burned out the finals in my transceiver with their tuning and transmitting. My interest in rhombics is now at a low ebb.

The long wire antenna has come into wider use lately. Recently, I heard an Alaskan operator describe a unique way of getting his long wire into a tree top. He fixed a string onto an arrow which he shot over the top of the tree and then teased up his insulated 400-foot long wire. His signal was tremendous.

My experience with my own long-wire antenna tuner, and a commercial model as well, has indicated good performance but poor in the area of tvi. I now use my long-wire antenna tuner on my bamboo fishing pole balcony antenna cut for 15 meter CW, with a short coaxial transmission line to the tuner. The swr is nearly 1:1 and I have no television interference. **Fig. 3** shows my bamboo fishing pole balcony antenna which uses the metal railing as a ground plane. I use the ground in the electric base plug in the shack for the antenna tuner ground.

miniaturizing

The ham antenna can be shortened by use of several different techniques. You can insert a coil to shorten the overall length, or use a capacitance hat. The

addition of coils causes some lumping of the antenna which will reduce the Q factor and narrow the bandwidth somewhat, but this is minimal if your tailoring is good. Of course, bringing the ends of the elements off at an angle will forshorten the necessary space, too. The long-wire tuner will work with one-quarter wavelength antenna with a good ground. During vacation I have used my bamboo fish pole with 40 feet of wire

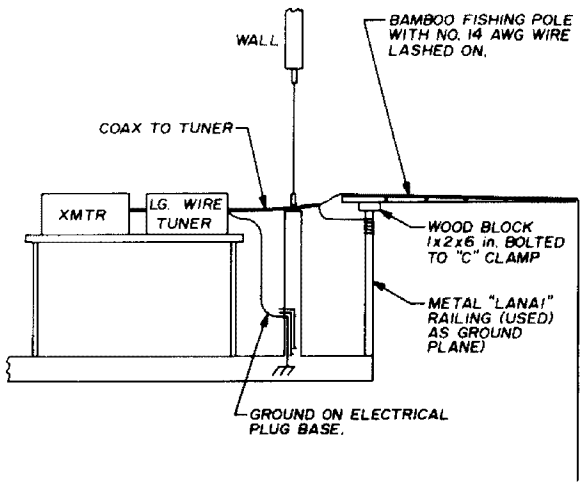


fig. 3. Fishing pole antenna with random length wire for portable operation on 10 through 40 meters.

over a balcony to work 10 through 40 meters with the tuner.

Hams are very ingenious when it comes to figuring out how to get a signal out. I know one amateur who put his miniature vertical slightly on its side in his attic and he works the world. Another friend of mine had a dipole with director and reflector in his attic. Balcony railings and drain pipes can sometimes work well as ground planes. For camouflage I have used pale blue auto spray paint on rope guys, insulators, baluns and coax.

Antenna tailoring has some very sound rules which I think should be followed. First of all, install the antenna high, in as clear an open area as possible. Avoid metal objects; this includes another antenna — such as a drooping dipole on a tower with metal or wire supports. Use rope where possible.

Resonate the antenna at the frequency you wish to use, cut the transmission line one-half wavelength long at your selected frequency and correct the input impedance to 50 ohms. Let the rest of the bands fall where they will if you have a multi-band antenna. Perfection on every band is impossible. Finally, learn to use the grid-dip meter and antenna bridge as these are your best antenna tailoring friends.

conclusions

A good antenna tailor will markedly improve performance on receiving and transmitting. Equally important is the fact that this improvement will often reduce or even eliminate radio, hi-fi, tv and telephone interference. Amateurs who are crowded closely together should include a bit of restraint in the use of power. A 100-watt signal will, with correct antenna tailoring, give you a lot of pleasure and relaxation. If you then add a linear often the entire neighborhood closes in. I have a friend who works the world and has one of the strongest signals from Honolulu. He uses only 500 watts. All of this is done in a large apartment unit packed with potential interference problems.

Don't think that tailoring means that you should build all of your own antennas. Most of us have neither the skill nor knowledge to compete with commercial antennas. Factory-made beams and multiband verticals are far more attractive and function far better than what I can build. The electronic engineers who build these antennas do a good job of setting resonant frequencies and matching impedance. Their testing is done under far better conditions than most of us have at our disposal and often, after carefully following factory specifications, you will find that the antenna resonates out of the ham bands. Your intelligent adjustments will bring the swr and directivity in line with what the direction sheets show. Currently, my antenna building handicraft is reserved for dipoles, long wires and tricky little antennas for limited space. However, I have found antenna tailoring to be a challenging and rewarding part of the art of amateur radio.

ham radio

four-element collinear antenna for 440 MHz

Complete construction details for a high-performance, omnidirectional antenna for 440 MHz that is built from easy-to-find materials and provides up to 6-dB gain over a dipole

Juan Rivera, WA6HTP, 1738 Berkeley Way, Berkeley, California 94703

Regardless of its design, a vertical omnidirectional gain antenna achieves its gain by the same process. It simply compresses the energy down on to the horizon where it is useful, and as a result, it wastes less radiating into the ground and up into the sky. Uhf antennas usually accomplish this by feeding several vertically mounted elements in phase so that the energy will combine itself into a pattern similar to that shown in **fig. 1.** These elements can all be fed in series, as Franklin, the originator of the collinear, did. This is the same system used by many commercial models (**fig. 2**). The elements can also be fed in parallel as I have done.

At first the series method seems much easier. A few inverted pieces of coax or a wire with some bends here and there and you're done. However, this approach has several big disadvantages. First of all, all the construction errors are cumulative. If you're slightly out of resonance at the

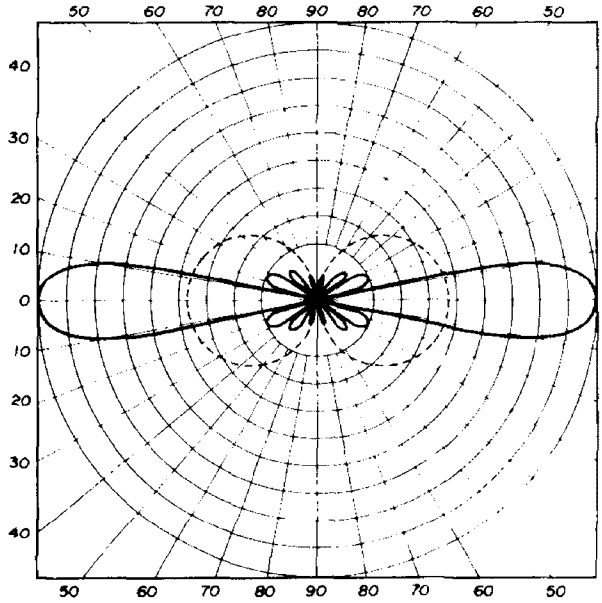


fig. 1. Vertical field strength of the four-element collinear (solid line) vs the dipole (dashed line). Collinear elements are positioned for omnidirectional pattern.

bottom, you can be sure that you're way out at the top. By its nature the series-fed collinear is a very touchy thing to get working properly and has very narrow band-width. Also, it must be supported by a low-loss pole or suspended from above. If it doesn't work when it's built, you have your hands full trying to find out why.

On the other hand, the parallel approach offers several advantages. First, all the elements exhibit 50-ohm input impedance and can be individually adjusted with a wattmeter. Secondly, the design is very rugged and the metal support pole can be grounded for lightning protection. In addition, the bandwidth of the antenna is at least 30 MHz and the elements can be positioned for either an omnidirectional or an offset pattern. Finally, it *will* work the first time you try it!

the dipole

Since the theory of the dipole is well known and covered in great detail in most antenna discussions, I won't dwell on it. The reason I selected the dipole over a folded dipole or a J-pole configuration is because of its simplicity and very wide bandwidth. It is constructed from 1/2-inch copper water pipe and 1/4-inch soft-drawn copper tubing. Both are excellent conductors and are readily available from most hardware stores.

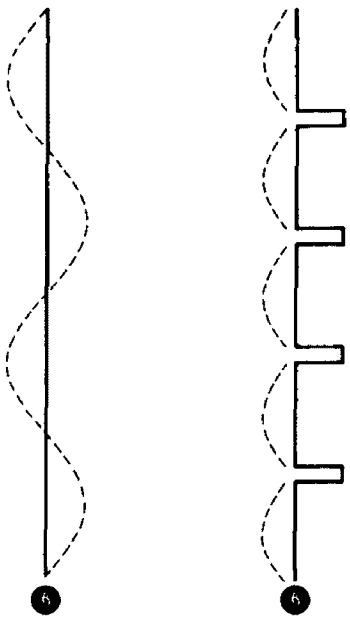


fig. 2. Franklin antennas. Natural standing-wave current distribution is shown in (A). Co-phased standing-wave current distribution is shown in (B).

As shown in fig. **3**, the impedance of the dipole increases from the center to a maximum of about 70 ohms at the ends. The exact impedance depends on the length-to-diameter ratio of the element as well as the distance from other objects.

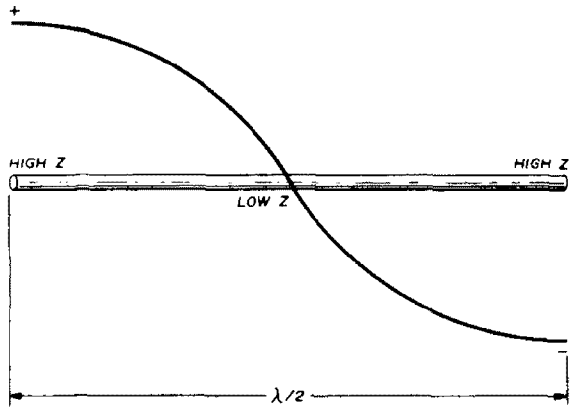


fig. 3. Half-wave dipole. Since there is no voltage to ground at the center, the element may be grounded to the support mast at this point.

At one point along either side a place will be found that will match perfectly into a 50-ohm transmission line. This is where the gamma rod is attached. Because of all the unknown variables, this point must be found by trial and error. With the antenna described here the gamma rod is attached 3-inches from the center.

To compensate for the inductive reactance of the gamma rod and to achieve a perfect impedance match, a series gamma capacitor is required. This consists of a length of number-12 Teflon-insulated wire fitted inside the end of the gamma rod (fig. **9**). A variable capacitor could be used but it would be hard to find one as inexpensive and weatherproof as the Teflon-insulated wire. Teflon was picked for its excellent electrical properties as well as for its resistance to the weather.

If Teflon is not available the center conductor of some RG-58/U coax will do, but the length will not be quite the same. If you use number-12 Teflon-insulated wire and follow the dimensions,

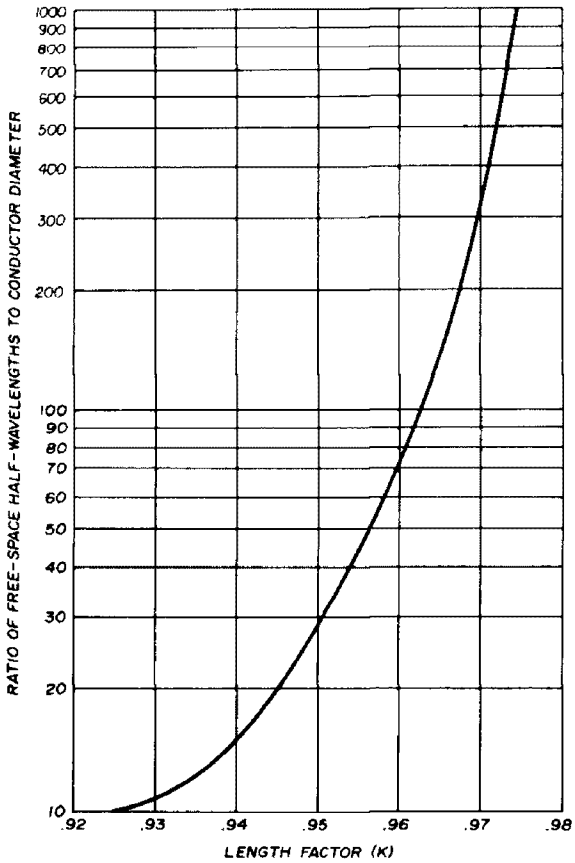


fig. 4. Effect of element diameter on the length of a half-wave dipole.

chances are that no tuning will be required.

As with the input impedance, dipole length is also affected by several variables. The approximate length is given by the formula

$$\text{length (inches)} = \frac{5905k}{\text{freq (MHz)}}$$

The k factor compensates for the shortening effect that increasing element diameter has on the length. This is plotted in fig. 4. With a length-to-diameter ratio of 24:1, k is equal to 0.9475. Therefore,

$$\text{length (inches)} = \frac{(5905)\ (0.9475)}{\text{freq (MHz)}} =$$

$$= \frac{5595}{\text{MHz}}$$

or a length of 12.7 inches at 440 MHz. Again, unfortunately, the graph of fig. 4 doesn't take into account the additional shortening effect of the support mast. The increased capacitance of the mast causes the actual resonant length to decrease by an additional 3½% (12¼ inches). This is mentioned mainly for those of you who wish to scale this basic design to another band. This information will get you close enough so that only a little trimming will be necessary.

matching harness

The function of the matching harness is to split the power evenly and deliver it to the elements in phase and at the proper impedance. Fig. 5 shows that, in a transmission line of even half wavelengths, the voltage and current are exactly the same at the load as at the input. This is true regardless of the vswr, assuming a lossless line. Therefore, it follows that the impedance is also the same at both ends.

The graph of fig. 5 also shows that for an odd 1/4-wavelength section the input and the load impedance are inverted. In other words, if at the input the voltage peaks and the current nulls, at the load the current would be at a peak and the voltage would null. This relationship is given by

$$\text{input impedance} = \frac{Z_o^2}{Z_r}$$

Where Z_o is the characteristic impedance of the line and Z_r is the load impedance (must be resistive).

Now, look at the harness schematic in fig. 6. Since each 50-ohm dipole is fed with a one-wavelength section the impedance at the input end is also 50 ohms. The four 50-ohm sections pair into two tees. Therefore, the impedance at the tees becomes 25 ohms. The tees are then

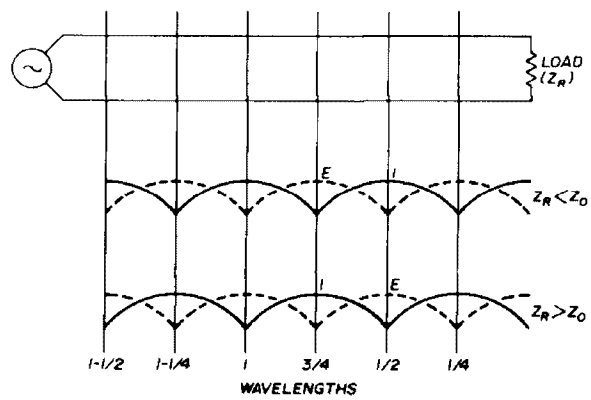


fig. 5. Standing waves on a transmission line terminated in a resistive load.

connected to a 1-1/4 wavelength section of transmission line. Therefore, from the above equation

$$\text{input impedance} = \frac{50\ \text{ohms}^2}{25\ \text{ohms}}$$

$$= \frac{2500}{25} = 100\ \text{ohms}$$

The two 5/4-wavelength lines are then connected through a tee, providing an input impedance of 50 ohms.

antenna spacing

With the dipoles and the harness out of the way only two things remain. The dipole to mounting pole spacing and the dipole to dipole spacing.

The horizontal field-strength patterns in fig. 7 were plotted using one of the dipoles mounted with a clamp to a pole on the roof of my house. The skewing of the patterns is due to the large relative

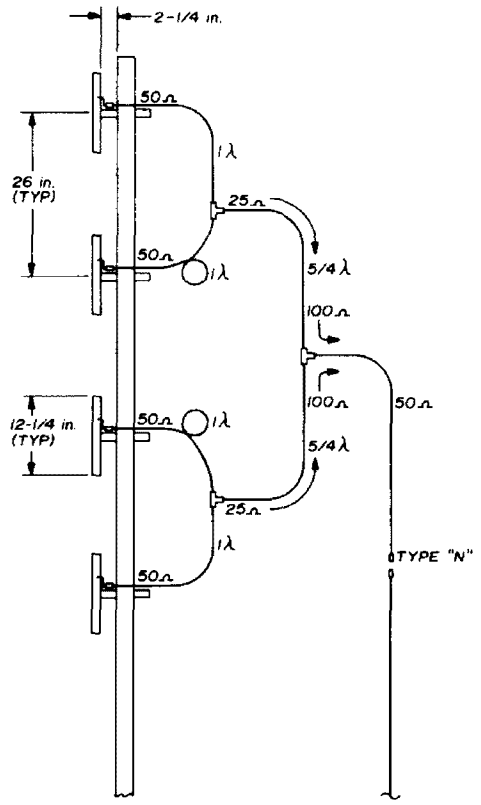


fig. 6. Layout of the matching harness for the four-element collinear. All sections are made from foam-filled Belden 8219 coaxial cable. The 1-wavelength sections are 20-15/16" long from the center of the tee to the solder terminal of the jack. The 5/4-wavelength sections are 26-11/64" long, measured between the centers of the tee connectors. All connectors are weatherproof BNC units (see parts list).

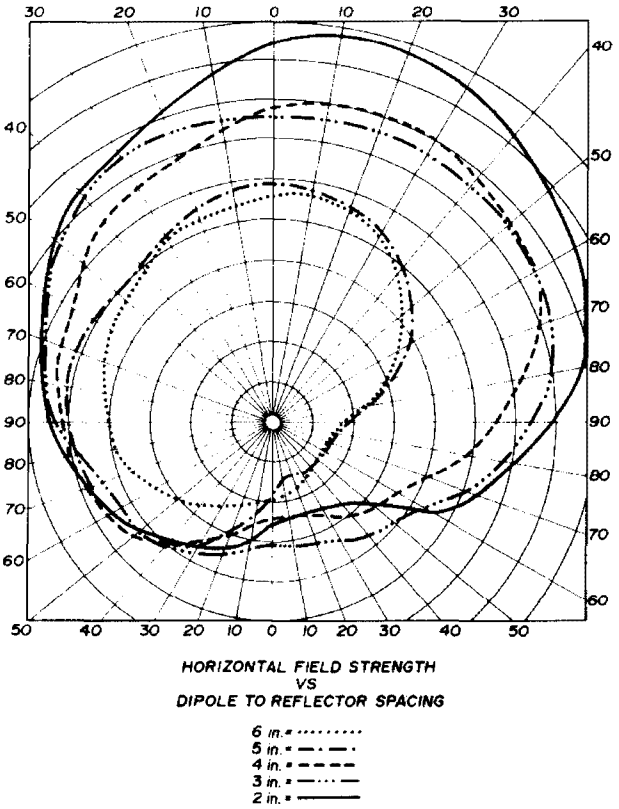


fig. 7. Field strength of single dipole element versus various dipole-to-reflector spacings.

size and close proximity of the clamp. I found that a spacing of 2- to 2¼-inches produced the best match and the highest gain. This corresponds to a spacing of about 0.11 to 0.12 wavelengths.

I didn't run any tests on vertical spacing, but research by others on similar antennas has shown that the maximum gain for a four-element array is achieved with a spacing of 0.97 wavelengths (26 inches at 440 MHz). This, incidentally, is the main reason for using the foam-filled coax. Belden 8219 coaxial cable has a velocity factor of 0.78 as opposed to only .659 for RG-58/U. Therefore, the length of harness made of 8219 cable will be correspondingly longer, thus allowing the elements to be spaced for maximum gain. If RG-58/U is used, the harness will not quite reach, and the spacing has to be reduced slightly. Since the difference is only a few inches, only a few tenths of a dB are lost, so don't worry if you can't find Belden 8219 coaxial cable.

construction

I would suggest that all the copper and

brass be polished before it is cut up, as this is much easier than doing it afterwards and will help to insure good solder joints. If you are going to use a gas stove or a torch be sure to do the actual soldering away from the flame. Otherwise, the rosin will burn and won't do its job.

Remember, at these frequencies the rf is concentrated in the outer few thousandths of an inch of conductor and a good solder joint is more important than ever. After soldering, cooling can be speeded up a bit by dripping a small amount of water on the end of the pipe *farthest* from the solder joint.

However, don't overdo it. Also, be very careful not to move the solder joint while it is cooling. A good solder joint will have a smooth appearance and a nice even fillet. If it's dull and grainy looking it was either moved before it solidified or was cooled too fast.

Dipole construction is simple and straightforward. A tubing cutter is a big time saver and does a very neat job. Other than that, no special tools are required. First, take the 1/2-inch pipe and cut it into eight pieces, each 5-3/4-inches long. In four of these pieces drill a 1/4-inch hole 3-inches from one end (fig. 9). From the remaining piece of pipe cut four pieces, each 4-1/2-inches long.

Next comes the 1/4-inch tubing for the gamma match. Make a 3/8-inch radius bend of 90-degrees and cut to size as shown in fig. 9. Be careful to leave at least 1/4-inch extra where the gamma rod

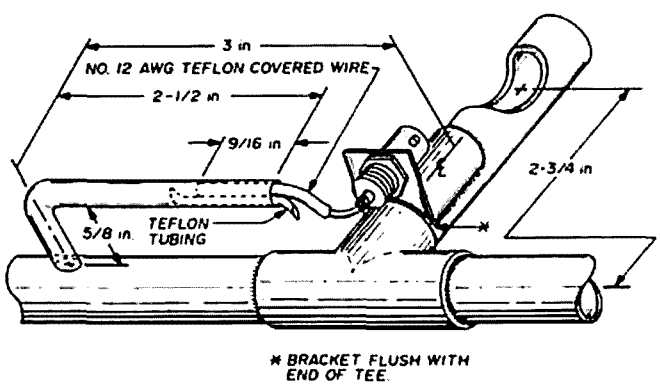
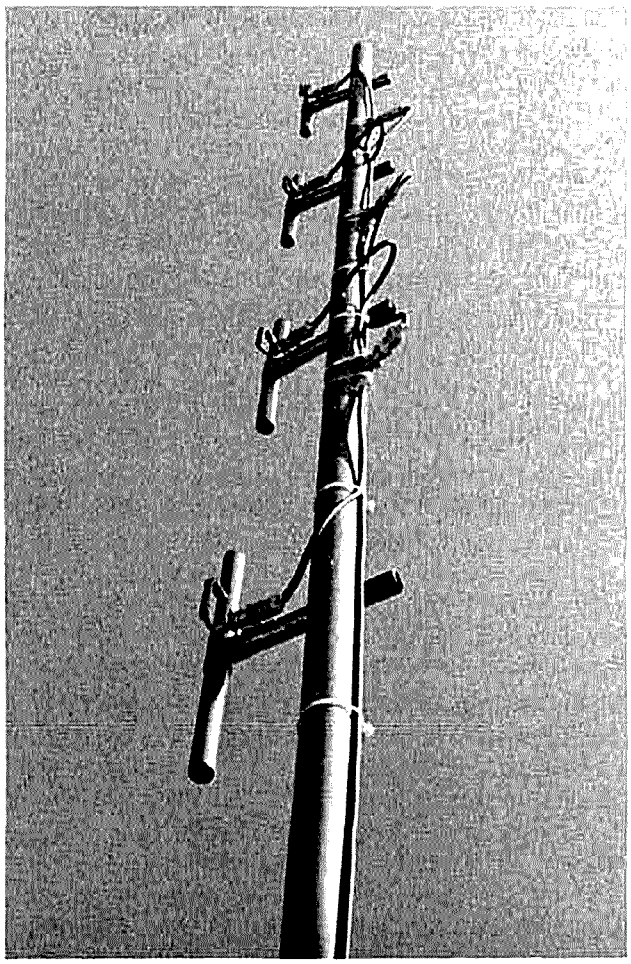


fig. 9. Construction of the gamma match. The Teflon tubing is used only to hold the wire firmly in place.



Collinear antenna with elements placed for skewed pattern. Gain in this configuration is approximately 9 dB. Note placement of harness away from elements.

mounts into the 1/2-inch element. The important thing here is to avoid kinks. After the 1/4-inch gamma rods have been bent, deburr the end that the wire will go into, using a small round file or a drill.

To make the support brackets cut a piece of copper or brass sheet 5/8-inch wide by 6-inches long and drill four 3/8-inch holes as shown in fig. 10. Cut and bend the brackets as shown, and then, using a vise or pliers, bend the side that is to be soldered to the pipe to conform to it.

Now, you are ready to begin soldering. First, solder the gamma rod to the pipe element. Pinch the end of the gamma rod a bit so you have a tight fit; then line it up. Hold the gamma rod and the element steady with a damp rag and start soldering. When it cools off, if it's a little out of line, it can easily be bent into shape. Next, assemble the element, temporarily

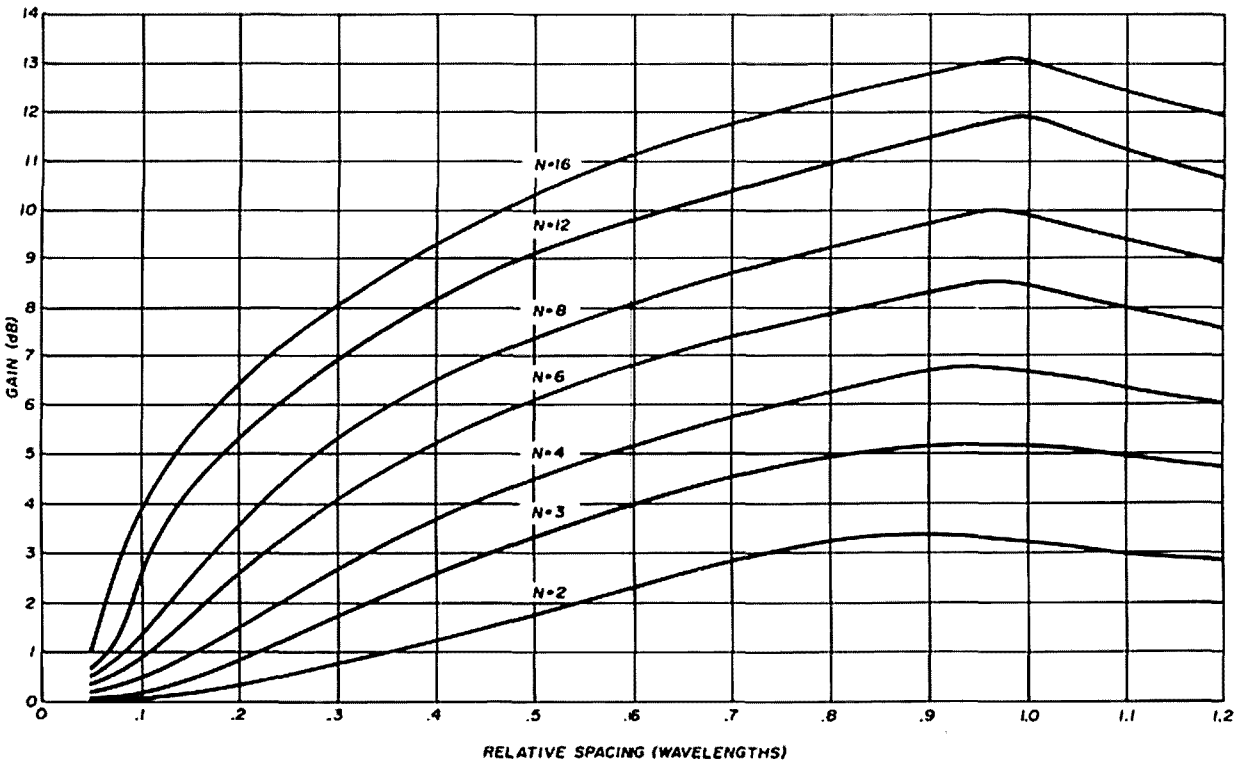


fig. 8. Gain to be expected from an omnidirection collinear array of dipole elements relative to the gain of a single element. N = number of elements.

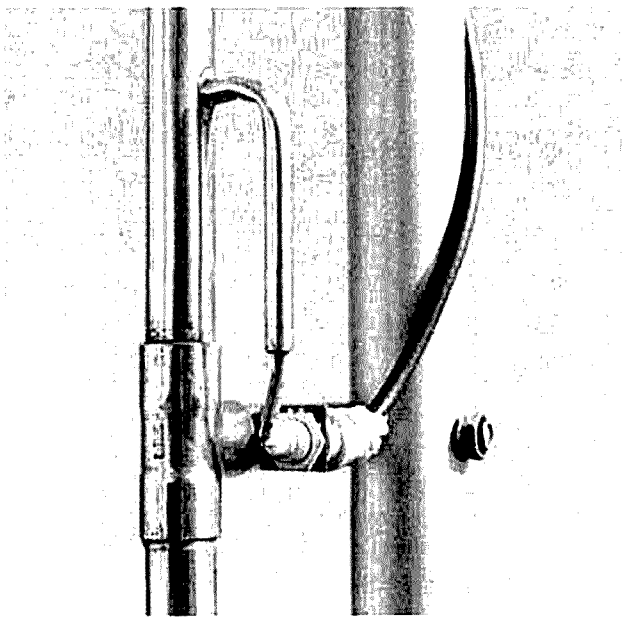
leaving off the connector and the wire.

With the dipole laying on your bench the bracket and the gamma rod should point straight up with the bracket flush with the tee. Drill a small hole through both the bracket and the pipe, and run a screw through them to hold everything steady while you're soldering (the screw can then be removed).

To keep the gamma rod from coming loose while you're soldering the bracket, hold the dipole with a damp rag at the gamma-rod end. Now, solder all the parts together.

When it cools you are ready to cut the mounting hole in the support arm (fig. 11). First, cut a V-shaped hole with a hack saw, being careful to keep the saw blade parallel with the element. Then file the hole to the same radius as the mounting pole. Check your progress often by fitting it to the mast. It's important for the spacing to be an even 2¼-inches the whole length and for the dipole to be parallel with the mast.

You might think that a simple clamp would be much easier, but a glance at fig. 7 shows what that does to the pattern. Now, give the whole thing a quick going over with the polish and install the connector and the wire. Make sure no strands are sticking out beyond the insulation that could short to the copper tubing. Leave a slight bend in the wire so that it can be adjusted a little each way from the 9/16-inch depth. Jam a piece of Teflon tubing in to hold the wire securely in place.



Construction of the dipole elements, showing the layout of the gamma match.

If you have access to a wattmeter you can now temporarily mount each element on the mast and tune for minimum reflected power. Keep everything, including your body, at least ten feet away when taking a reading. Tape the coax out of the way behind the pole and run it down to the ground. Only a slight adjustment of the wire depth, if any, should be required (with 50-watts input the reflected power should be much less than a watt).

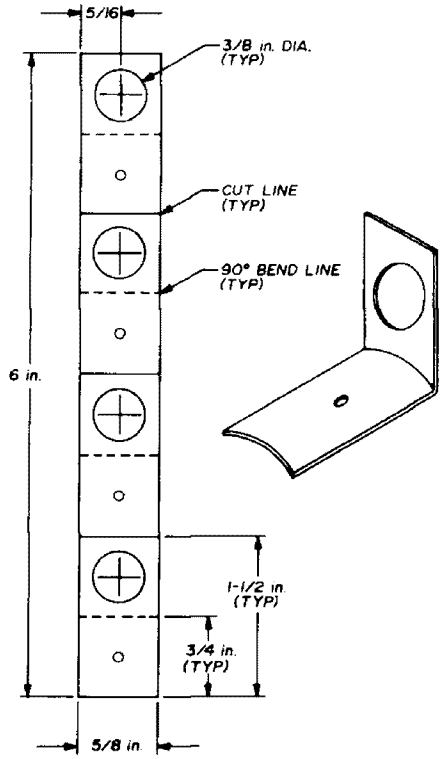


fig. 10. Construction of the mounting bracket. Material is thin brass sheet.

When each element has been tuned remove the rosin and polish with a solvent such as undiluted alcohol, tape over the mating end of the BNC connector and you're ready to put on a coat of epoxy. I sprayed the epoxy on using an aerosol power pack but it can also be brushed on. Cover everything thoroughly. If it's not covered it will corrode, and corrosion is a lousy conductor! When you are done hang the elements up to dry.

harness construction

The construction of the harness is very simple. The lengths are found from

$$\text{wavelength (inches)} = \frac{11800 \text{ (velocity factor)}}{\text{freq (MHz)}}$$

Remember that the connectors are considered to be part of the coax cable and must be included in your measurements **(fig. 6)**. I found it easiest to stick a length of masking tape down on my work bench and mark the length out on it. Install a BNC connector on one end of the coax, connect it to a tee and then measure from the center of the tee. I intentionally made my cables a little long and then figured out how much shortening was necessary to make them the right length.

When you have finished all the matching sections assemble the harness and check for continuity between all the center pins. Now, liberally coat all the places that water could ever conceivably enter with silicon sealant.

final assembly

After everything is dry all that is necessary is to mount everything on the mast. If you use the Belden 8219 coax you can use the maximum gain spacing of 0.97-wavelength. If you use RG-58/U you will have to mount the elements slightly closer together. Leave a foot between the top of the mast and the top of the first dipole. Line it up and drill a hole through the pole and the pipe **(fig. 11)**.

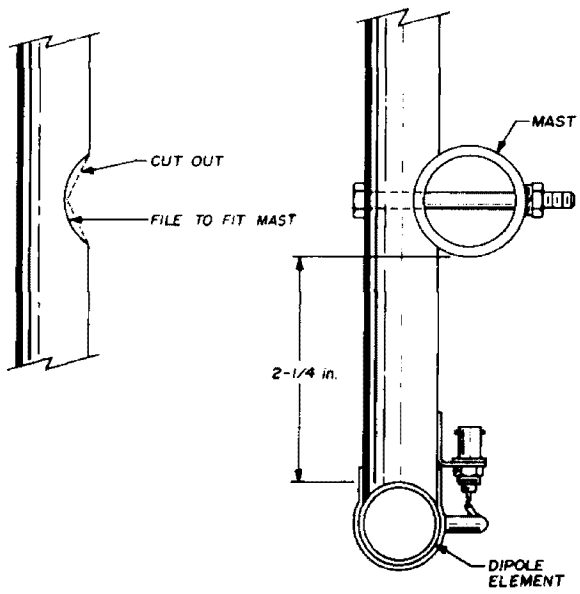


fig. 11. To mount each individual dipole on the mast a small circular section must be cut out of the support pipe.

When you tighten the mounting bolt don't get carried away. The copper pipe deforms quite easily. Be sure to use a lock washer or a self-locking nut. Then run the cable around to the back, away from the element, and secure it with a plastic cable tie or tape.

Next, decide what kind of a pattern you want. The omnidirectional pattern has about 6 dB gain over a dipole and the offset pattern, about 9 dB toward the front (**fig. 12**). If you want the offset pattern all you have to do is mount all the elements facing the same way. If you want the omni pattern point them 90-degrees apart. Mount the bottom element as far down the mast as it will easily go if you use RG-58/U or 78-inches if you use Belden 8219 coax. Don't cheat and use a BNC right-angle connector as that would change the length of the line and mess it up. Next, mount the remaining two elements evenly between the others and secure the harness in place behind the dipoles.

conclusion

If some of the materials listed are not available, don't be put off. Improvise. The BNC connectors were selected because they are *improved* and weatherproof. However, if they are all sealed with silicon sealant, as I have suggested, the older type BNC connectors should work just fine. In fact, if a low-loss weatherproof connection can be devised, no connectors would be needed at all. This would save considerable money and might allow the harness to fit inside the mast out of the weather.

The pool skimmer pole I used for the mast was selected only because of its low cost and availability in California. It can be replaced by almost anything. I hope that you have as much fun building this antenna as I had — you will be pleased with the results.

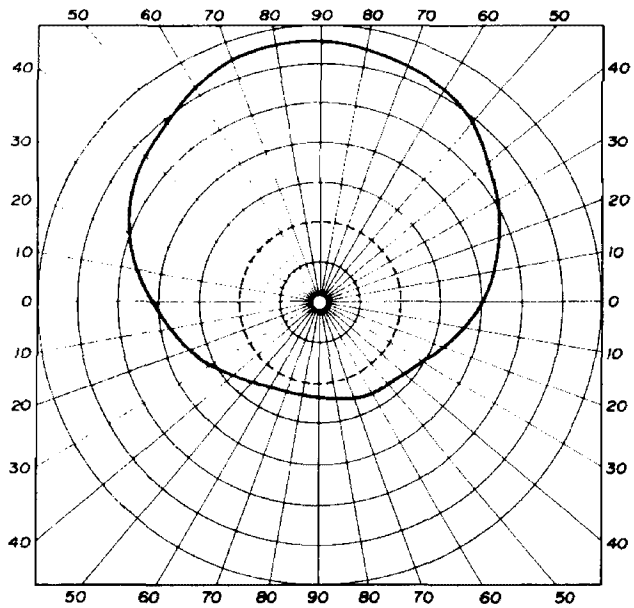


fig. 12. Horizontal field strength of the four-element collinear with all elements facing front. Gain in this configuration is approximately 9 dB. Single dipole is shown for reference (dashed line).

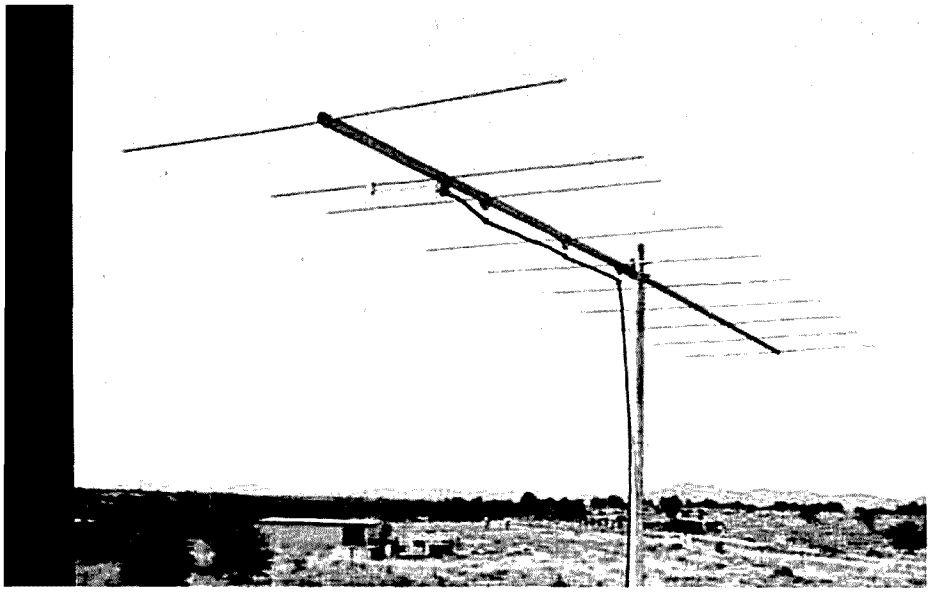
table 1. Parts list for the four-element 450-MHz collinear antenna.

8 feet	½" hard-drawn copper pipe
2 feet	¼" soft-drawn copper tubing
20 feet	Belden 8219 foam-filled coaxial cable
1 foot	1/16" Teflon tubing
1 foot	number-12 Teflon-insulated wire
1 pint	clear epoxy paint
1 tube	silicon sealant
1 can	Brasso metal polish
4 sets	stainless-steel hardware (2" x ¼" bolts, nuts and washers)
4	½" copper tees
1	brass sheet, 6" long, 5/8" wide
1	12- or 17-foot single-section aluminum swimming pool skimmer pole (or equivalent for mast)
13	UG-88E/U BNC cable-mounted male connectors
3	UG-274B/U BNC tee connector
4	UG-1094A/U BNC female bulkhead mount connectors
1	UG-556B/U type-N female cable connector

references

1. D. Blakeslee, W1KLK, *The Radio Amateur's Handbook,* 49th edition, A.R.R.L., Newington, Connecticut, 1972.
2. E. Tilton, W1HDQ, *The Radio Amateur's VHF Manual,* A.R.R.L., Newington, Connecticut, 1972.
3. H. Jasik, *Antenna Engineering Handbook,* McGraw-Hill, New York, 1961.

ham radio



how to design gamma-matching networks

A complete procedure for designing gamma and tee antenna-matching networks

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When the driving-point impedance of thin linear antenna systems is known, these systems can generally be gamma- or T-matched to a transmission line with excellent results. The gamma-match is used to match an unbalanced coaxial line to either a monopole or a dipole driven element, and the T-match is used to match a two-wire balanced line to a dipole driven element.

I will first discuss the gamma matching procedure, and will follow that with the T-match, showing that it is simply an extension of the technique used to design a gamma match.

gamma match

The layout of the gamma-match, minus the capacitor, is one-half that of a T-match, as shown in fig. 1. In this illustration L is the length of one-half of the driven dipole (total length of the driven monopole) and is often near one-quarter wavelength long. The length of the gamma rod, which is parallel to the driven element, is labeled l_r. The center-to-center spacing between the gamma rod and driven ele-

ment is S, and Z_i is the input impedance to the gamma network, including the driving-point impedance.

When the element and gamma-rod diameters and spacings are *much* less than the carrier wavelength at which the system will be used, the geometry of **fig. 1** is analogous to the electrical circuit shown in **fig. 2.**[1,2,3] In **fig. 2** the input impedance, Z_i, is equal to

$$Z_i = \frac{(H_z Z_a)(X_s)}{H_z Z_a + X_s} \text{ ohms} \qquad (1)$$

where Z_a is one-half the dipole driving-point impedance (or monopole driving impedance), X_s is the reactance of the gamma rod and H_z is the antenna impedance step-up ratio.

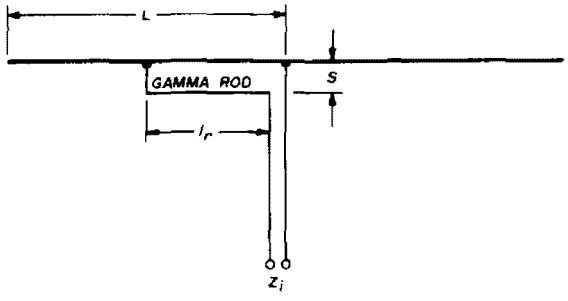


fig. 1. The basic gamma match. L is one-half the length of the driven dipole.

The reactance of the gamma rod X_s, is given by

$$X_s = jZ_o \tan k l_\Gamma \text{ ohms} \qquad (2)$$

where $Z_o = 60 \cosh^{-1}\left(\frac{4S^2 - D^2 - d^2}{2Dd}\right)$ ohms

in air* and $k = 2\pi/\lambda$ radians per meter. The reactance, X_s, is positive when the gamma rod electrical length, given by kl_Γ, is less than $\pi/2$ radians, as will be assumed here.

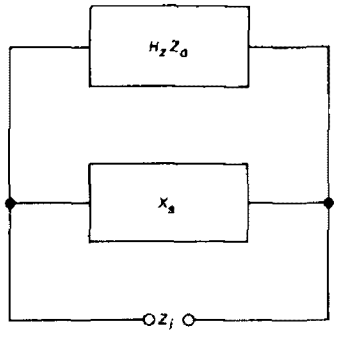


fig. 2. Equivalent circuit to the gamma-matching system shown in fig. 1. Za is one-half the dipole input impedance, Hz is the impedance step-up ratio and Xs is the reactance of the gamma rod.

The formula for the numeric antenna impedance step-up ratio, H_z, is

$$H_z = \left[1 + \frac{\cosh^{-1}\left(\frac{4S^2 - D^2 + d^2}{4Sd}\right)}{\cosh^{-1}\left(\frac{4S^2 + D^2 - d^2}{4SD}\right)}\right]^2 \qquad (3)$$

where D is the diameter of the driven element, d is the diameter of the gamma rod and S is the spacing between the two. This factor is plotted in normalized form in **fig. 3** for use in gamma-match designs. In plotting this graph it was assumed that H_z is at least 4:1, a realistic assumption in most amateur applications.

When the antenna is a self-resonant folded monopole the length of the gamma rod approaches $\pi/2$ electrical radians and X_s becomes much larger than $H_z Z_a$. In this case **eq. 1** is reduced to

$$Z_i = H_z Z_a \text{ ohms} \qquad (4)$$

In this case Z_a is not highly resistive and a low vswr at the junction of the folding element and the coaxial line may not be possible.

The complex input impedance, Z_i, of **eq. 1** may be written as a magnitude M and phase angle as given here

$$Z_i = M \angle \psi \qquad (5)$$

*I derived the characteristic impedance $\sqrt{L/C}$ for the high-frequency lossless line in air from the farads/meter length given on page 78 of *Static and Dynamic Electricity*, W. Smythe, McGraw-Hill Book Company, Inc., 1968, along with an approach similar to that given on page 210 of reference 13. The same solution appears on page 22-23 of *Reference Data for Radio Engineers*, Fifth Edition, ITT Staff, Howard W. Sams & Company, 1969.

where

$$M = \frac{X_s H_z \sqrt{R_a^2 X_a^2}}{\sqrt{(H_z R_a)^2 + (X_s + H_z X_a)^2}} \text{ ohms}$$

$$\psi = \tan^{-1}\left[\frac{H_z (R_a^2 + X_a^2)}{X_s R_a} + \frac{X_a}{R_a}\right] \text{degrees}$$

To match a lossless high-frequency transmission-line characteristic resistance, R_o,

$$M \cos \psi = R_o \text{ ohms}$$

The lumped reactance, X_Γ, added in series with Z_i to tune out or cancel the reactive component of Z_i can be determined from

$$M \sin \psi = -X_\Gamma \text{ ohms}$$

For **eq. 5** to have the correct magnitude and phase angle,

$$X_s = H_z \left\{ \frac{R_o X_a + \sqrt{(R_o X_a)^2 + R_o [H_z R_a - R_o] [(R_a)^2 + (X_a)^2]}}{H_z R_a - R_o} \right\} \text{ohms} \quad \textbf{(6)}$$

From which

$$X_\Gamma = -R_o \tan \psi$$

$$= -R_o \left\{ \frac{H_z [(R_a)^2 + (X_a)^2]}{X_s R_a} + \frac{X_a}{R_a} \right\} \quad \textbf{(7)}$$

The electrical length of the gamma rod (see **eq. 2**), $k l_\Gamma$, is equal to

$$k l_\Gamma = \frac{2\pi l_\Gamma}{\lambda} \cong \frac{2\pi f_{MHz} l_\Gamma (360)}{(0.956)(300)(39.37)(2\pi)}$$

$$\cong 0.03188 f_{MHz} l_\Gamma \text{ degrees} \quad \textbf{(8)}$$

The quantity l_Γ is the length of the gamma rod in inches and the 0.956 coefficient is an average relative velocity along the gamma rod and radiating elements.[4]

As I was developing **eqs. 6** and **7** two ratios often appeared in my calculations. I have called these T and Q:

$$T = \frac{H_z}{Z_o} \text{ ohms}^{-1} \quad \textbf{(9)}$$

$$Q = \frac{X_s}{H_z} = \frac{\tan k l_\Gamma}{T} \text{ ohms} \quad \textbf{(10)}$$

These two ratios are useful observations because H_z and Z_o are a function of S/D D/d. That is, when H_z is determined, T can be found, and a solution for Q eliminates a separate calculation for Z_o. **Fig. 4** is a normalized graph of T vs S/D and D/d, and **fig. 5** is a plot of $k l_\Gamma$ in degrees vs Q and T. These graphs are extremely helpful when designing a practical gamma matching system.

When **eq. 10** is substituted into **eq. 6**, a solution for Q is a positive discriminant root of a quadratic equation as follows,

$$Q = A + \sqrt{A^2 + B} \text{ ohms} \quad \textbf{(11)}$$

Where

$$A = \frac{R_o X_a}{H_z R_a - R_o} \text{ ohms}$$

$$B = \frac{R_o [(R_a)^2 + (X_a)^2]}{H_z R_a - R_o} \text{ ohms}^2$$

When **eq. 10** is used with **eq. 7** a solution for the reactance of the gamma capacitor can be obtained

$$X_\Gamma = -\frac{1}{2\pi f C_\Gamma} = -[E + F] \text{ ohms} \quad \textbf{(12)}$$

This equation may be rearranged to provide the value of the gamma capacitor in pF

$$C_\Gamma = \frac{10^6}{2\pi [E + F] f_{MHz}} \text{ pF} \quad \textbf{(13)}$$

Where

$$E = \frac{R_o}{R_a}\left[\frac{(R_a)^2 + (X_a)^2}{Q}\right] \text{ohms}$$

$$F = \frac{R_o}{R_a} X_a \text{ ohms}$$

In **eqs. 12** and **13** the sum of E and F is positive. Otherwise, the phase angle is in the fourth quadrant and the gamma-matching capacitor C_Γ cannot be used. For the reactance of the gamma match, X_s, to remain positive, the electrical length of the gamma rod, $k l_\Gamma$, must be less than 90 electrical degrees and **eqs. 6** and **11** have the restriction

$$H_z > \frac{R_o}{R_a} \quad \textbf{(14)}$$

This equation is very important to broadband array operation. That is, R_a may change considerably with frequency changes and H_z should be large enough to assure the restriction given by **eq. 14**. Furthermore, vswr readings over a broad band reflect the combined characteristics of the antenna and the matching network, and a well designed gamma- or T-network will follow the inverse behavior of the antenna driving-point impedance.

Large values of H_z imply small values of S together with large wire diameter ratios, D/d. At the same time, the electrical length of the gamma rod, kl_Γ, should be greater than about 15 electrical degrees to minimize ohmic circuit losses as well as to maximize antenna system bandwidth. However, large values of H_z along with large design values of kl_Γ can lead to construction errors and/or instability in H_z, Z_o, T and Q. When the driven element diameter is not very large an H_z of more than 20 to 30 may not be practical.

When the driven element or parasitic elements of an array are arbitrarily pruned, element length and/or spacing, to obtain a resistive driving-point impedance, the forward gain may be reduced. There are many combinations of R_a, X_a, H_z, Z_o and l_Γ which satisfy **eq. 6**. Also, many *combinations* of D, d, S and C which are correct solutions, but not necessarily best.

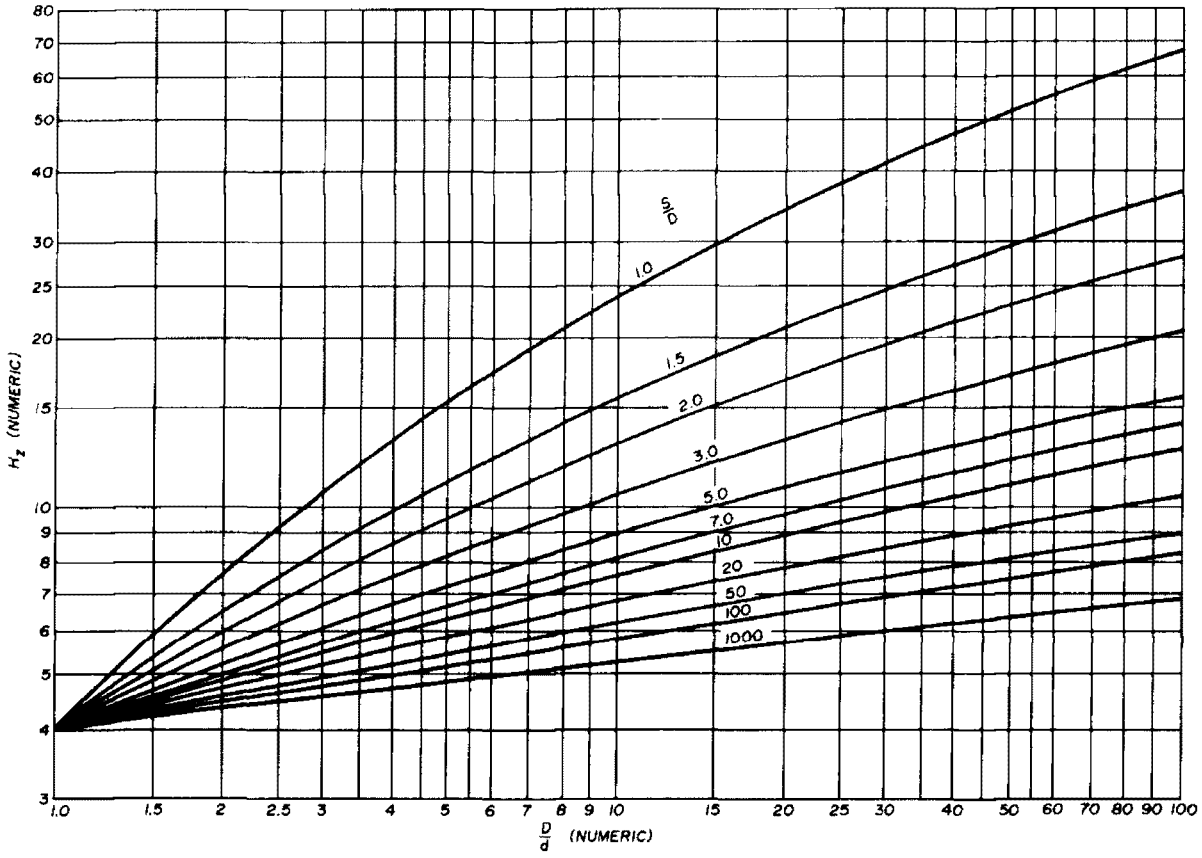


fig. 3. Antenna impedance step-up ratio, Hz, as a function of element diameters and spacing.

For example, if it is desirable to reduce the length of the gamma rod, **eq. 10** shows that Q will be reduced (T is inversely related). Looking further, **eq. 6** shows that a reduction in Q requires an increase in H_z, and **eq. 2** shows that an increase in H_z requires an increase in D/d or a decrease in S/D (or combination of both). From **eq. 3** it can be seen that an increase in D/d increases Z_o, while a decrease in S/D decreases Z_o. The ambiguity here is eliminated by plotting T, given in **eq. 9** and plotted in **fig. 4**.

practical gamma-match

Some time ago I obtained a homebrew 10-element, two-meter Yagi-Uda type in-line end-fire array with somewhat unusual spacings and element lengths. Measurements with the antenna's dissimilar diameter folded-dipole driven element connected to 300ohm open-wire line indicated that its best overall performance was near lower end of the 144-MHz band. An analysis of element lengths and spacing confirmed this observation, and calculations indicated that best performance was at about 144.6 MHz.

Since I wanted to use the antenna at a somewhat higher frequency, 145.4 MHz, I discarded the folding element (which transfers *any* antenna driving-point impedance, real or complex, to the transmission line) and shortened all the elements. I did not change element spacing or boom length, so performance was compromised somewhat. However, calculations indicated that the array's performance was much improved over the 145.35- to 146.25-MHz segment of the band. The free-space driving-point impedance, Z_a', is slightly inductive, about 14 + j3 ohms at 145.4 MHz where I'll be using the antenna most of the time.

To use the previously discussed gamma-match design equations, the free-space input impedance, Z_a', must be divided by 2. Therefore, Z_a = 7 + j1.5 ohms. Since I planned to use 50-ohm polyfoam RG-8/U with the antenna, R_o = 50 ohms. The diameter of the driven element is 0.375 inch and the diameter of the gamma rod, which is made from number-12 wire, is 0.0808 inch. The spacing between the elements, S, is approximately 0.6029 inch (0.375-inch long insulator plus one-half the diameter of the driven element, plus one-half the diameter of the gamma rod).

With these figures, the gamma-match design procedure can begin by finding two simple ratios:

$$\frac{D}{d} = 4.64$$

$$\frac{S}{D} = 1.61$$

When these two ratios have been found, the antenna impedance step-up ratio, H_z, can be found from **fig. 3**. In this case, H_z is equal to approximately 11.0. The quantity T can be determined with the help of fig. 4, approximately 0.046 ohm^{-1} in this case. Now, calculate A and B and use those quantities to find Q:

$$A = \frac{R_o X_a}{H_z R_a - R_o} \cong \frac{+75}{27} \cong +2.77 \text{ ohms}$$

$$B = \frac{R_o [(R_a)^2 + (X_a)^2]}{H_z R_a - R_o}$$

$$\cong \frac{2562.5}{27} \cong 94.9 \text{ ohms}^2$$

$$Q = +A + \sqrt{A^2 + B}$$

$$\cong +2.77 + \sqrt{102.57} \cong 12.9 \text{ ohms}$$

With the quantities Q and T known, the electrical length of the gamma rod, kl_Γ, in degrees can be determined with the help of **fig. 5**. In this case the gamma rod is approximately 31.0 electrical degrees long. This can be converted into inches by rearranging **eq. 8**.

$$l_\Gamma \cong \frac{kl_\Gamma}{0.03188 f_{MHz}} \cong \frac{31}{0.03188(145.4)}$$

$$\cong 6.69 \text{ inches}$$

To find the required gamma capacitance, calculate the quantities E and F from **eq. 13**.

$$E = \frac{R_o}{R_a}\left[\frac{(R_a)^2 + (X_a)^2}{Q}\right]$$

$$\cong \frac{2562.5}{(7)(12.9)} \cong 28.38 \text{ ohms}$$

$$F = \frac{R_o}{R_a} X_a \cong \frac{50(+1.5)}{7} \cong 10.7 \text{ ohms}$$

$$C_\Gamma = \frac{1{,}000{,}000}{2\pi [E + F] f_{MHz}}$$

$$\cong \frac{1{,}000{,}000}{35{,}700} \cong 28 \text{ pF}$$

Thus, the gamma-matching network consists of a gamma rod made from number-12 wire, approximately 6.69 inches long with an axial center-to-center spacing of 0.6029 inch from the *parallel* driven element. The gamma capacitor is about 28 pF, which can be provided by the 7- to 45-pF variable I happen to have in my junkbox.

This same basic design procedure can also be used if the driving-point impedance is slightly capacitive. For example, if my two-meter array driving-point impedance had been 14 - j3 ohms at 145.4 MH_z, Z_a for design purposes would be 7 - j1.5 ohms. The values for D/d, S/D, H_z, T, A and B would be the same as for the previous case, although the A quantity would have a negative sign. This changes the magnitude of Q to 7.36 ohms. Referring to fig. 5, the length of the gamma rod is approximately 20.0 electrical degrees, and its length is

$$l_\Gamma \cong \frac{k l_\Gamma}{0.03188 f_{MHz}}$$

$$\cong \frac{20}{0.03188(145.4)} \cong 4.31 \text{ inches}$$

Now, calculate the quantities E and F and determine the value of the gamma capacitor:

$$E = \frac{R_o}{R_a}\left[\frac{(R_a)^2 + (X_a)^2}{Q}\right]$$

$$\cong \frac{2562.5}{(7)\ (7.36)} \cong 49.73 \text{ ohms}$$

$$F = \frac{R_o}{R_a} X_a \cong \frac{50(-1.5)}{7} = +10.7 \text{ ohms}$$

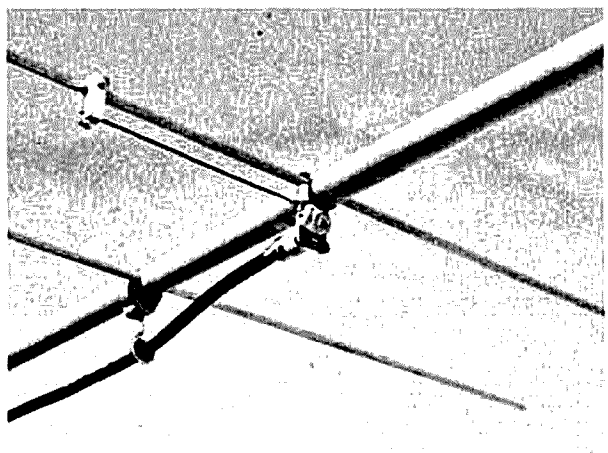
$$C_\Gamma = \frac{1{,}000{,}000}{2\pi\ [E + F]\ f_{MHz}}$$

$$\cong \frac{1{,}000{,}000}{35{,}700} \cong 28 \text{ pF}$$

Thus, when the gamma rod is shortened from 6.69 inches to about 4.31 inches, and no other changes are made, the gamma network will properly terminate the 50-ohm coaxial feedline into an antenna impedance of 14 - j3 ohms.

testing the design

Using an adjustable gamma rod 7.5 inches long and a 7- to 45-pF variable gamma capacitor, I assembled the gamma match to the antenna. When the antenna was erected horizontally, 18-feet above the ground, with the gamma rod adjusted to 6.75-inches long and the gamma capacitor set to approximately 25 pF, the vswr was as follows:



Gamma-matching system for the 144-HMz Yagi beam antenna.

146.3 MHz 1.38:1 vswr
145.4 MHz 1.03:1 vswr
144.9 MHz 1.16:1 vswr

As can be seen, at the design frequency of 145.4 MHz the standing-wave ratio is extremely small.

Calculations indicated that the driving-point impedance of this modified homebrew array becomes slightly capacitive at both 144.9 and 146.3 MHz. Therefore, I began to shorten the length of the gamma rod and increase the gamma capacitance to see if I could obtain a *constant* vswr. With a gamma rod length of 5.5625 inches and the gamma capacitor set to 35 pF I obtained what I was looking for:

146.3 MHz 1.27:1 vswr
145.4 MHz 1.28:1 vswr
144.9 MHz 1.27:1 vswr

I then twisted the antenna boom 90° so that the antenna was vertically polarized on the *metal* mast. Without touching the gamma adjustments, I slipped the boom back and forth until I obtained a minimum vswr of 1.3:1 at 145.4 MHz.

The boom was locked at this point, and both the gamma rod and gamma capacitor were adjusted (in that order) to obtain a minimum vswr of 1.09:1.

I then started to raise the height of the array by increasing the height of the mast. When the array was about two feet higher (13 quarter-wavelengths above ground) the vswr was maximum at 1.6:1. Above this height the vswr began to decrease. I expected maximum mast coupling to occur at odd quarter-wavelengths of mast, and this behavior confirms it.

I locked the antenna at 21 feet above the ground (I wasn't able to reach the gamma capacitor at greater heights) and adjusted both the gamma capacitor and the gamma rod. I then pushed the antenna up to slightly more than 25 feet above ground, the maximum height of the mast, and made the vswr measurements. After several adjustments at the 21-foot level, at the 25-foot height, with a gamma capacitance of 30 pF and a gamma rod length of 5.875 inches, I obtained the following vswr measurements:

146.3 MHz 1.10:1 vswr
145.4 MHz 1.02:1 vswr
144.9 MHz 1.04:1 vswr

performance

Essentially all of the two-meter stations in Tucson, about 35 miles from my station, are vertically polarized, and my propagation is by diffraction over the north end of the Santa Rita mountain range. In running tests with K7RMH in Tucson, I was delighted to find that my British dual-eight skeleton slot antenna, which is 3 to 4 feet higher than the homebrew array, was only about 3 dB better with the same power input. That is, with one-watt rf input, at K7RMH's station I was S9 +30 dB with the homebrew array and S9 +33 dB with the skeleton slot.

Upon further testing with a two-position coaxial switch in the feedline for rapid switching between the line tuner and the RG-8/U coaxial cable, I discovered that the coaxial switch generates a vswr of 1.5:1 at the transmitter. This results in

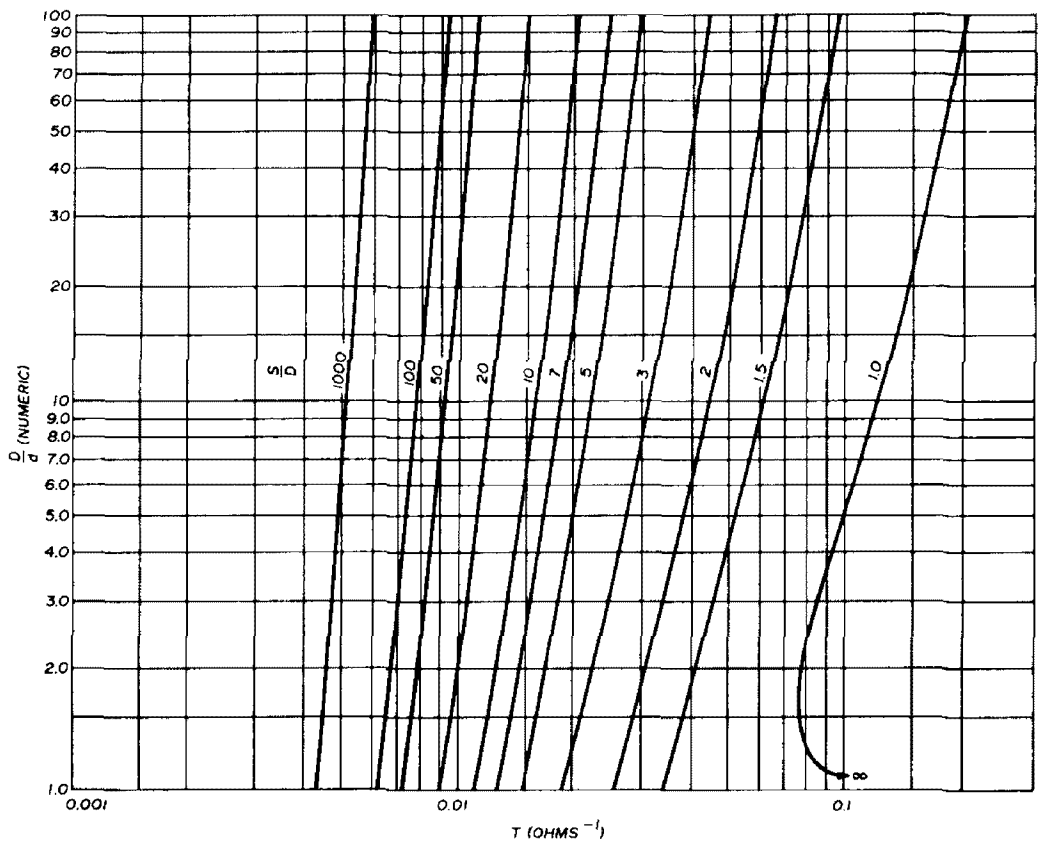


fig. 4. Ratio T (Hz/Zo) as a function of gamma-matching element diameter and spacing.

about a 1.0 dB loss at K7RMH's station which suggests that the K7RMH signal readings are intensity, *not* power, sensitive. This indicates that a change of 3 dB at K7RMH is a *power* change of 6 dB.

Calculations indicate that the gain of the skeleton slot is about 4 dB better than that of the homebrew array. I am using about 35-feet of 300-ohm open-wire feedline to the slot antenna, and 50-feet of RG-8/U polyfoam coax to the homebrew array. Assuming that 1 dB more is lost in the coaxial line than in the open-wire feeder, and that the height gain of the slot antenna is 1 dB better, the test results check out very well.

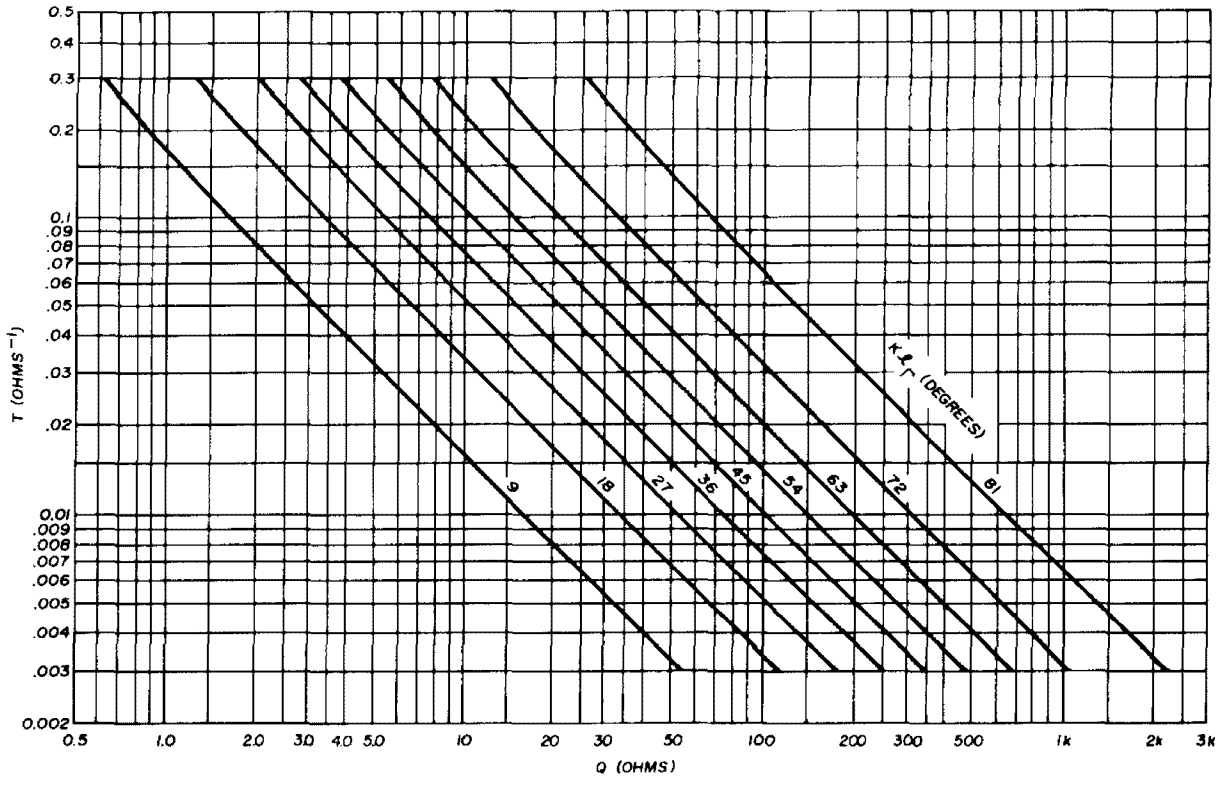


fig. 5. Length of the gamma rod in electrical degrees vs the quantities T and Q. This graph is very helpful when designing practical gamma-matching systems.

impedance measurements

A problem many amateurs face when designing their own antenna is measuring the driving-point impedance. Often, they do not have the necessary equipment or the technique is long and drawn out, and provides erroneous results. This seems to be underscored by an article in the July, 1972, *IEEE Trans. on Antennas and Propagation* calling for papers on antenna measurements.

It may be that the recent article by W3TQM which described how to determine antenna impedance by direct swr measurements may prove popular with amateurs.[5] I hope to try this procedure sometime to confirm my calculations on the homebrew two-meter array to see how it works out.

Another possibility is the simple complex impedance bridge described by W2CTK.[6] When compared to a commercial RX bridge the accuracy of the W2CTK bridge is quite good. A complete discussion on the use of this instrument appears elsewhere in this issue.

In the absence of impedance measurements, the amateur must generally resort to existing published material which provides the driving-point impedance for various antenna configurations. However, with different (unscaled) antenna layouts, the magnitude and phase angle of the driving-point impedance will be different.

However, a review of some of the

available material suggests that, most of the time, the resistive, R_a, and reactive, X_a, portions of the antenna driving-point impedance will fall within the following limits:

$$+3 \leqslant R_a \leqslant +175 \text{ ohms}$$

$$-70 \leqslant X_a \leqslant +55 \text{ ohms}$$

For estimates of the driving-point impedance, consult references 1, 2, 4, 7, 8, 9, 10, 11, 12, 13 and 14. Computer programs have been developed for calculating the driving-point impedance vs the layout of the antenna, but unfortunately, these programs are generally not available to amateurs.

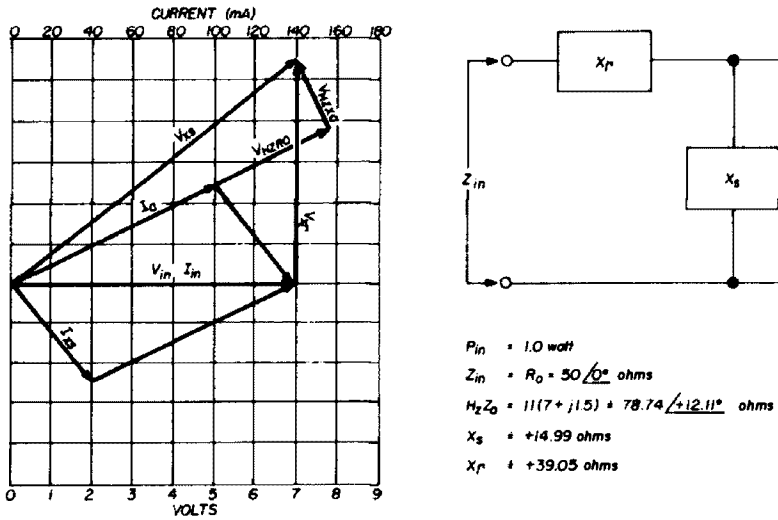


fig. 6. Voltage and current distribution of the gamma-matching network designed to match the 50-ohm coaxial line to the 10-element beam at 145.4 MHz (Za = 7 + j1.5 ohms). Voltage and current values are based on power input of 1 watt.

another design approach

When H_z can have a value of 4.0, **fig. 3** suggests that the S/D ratio can be any value when D/d = 1.0. In this case a decision as to what value of S/D to use can depend upon a *desired* kl_Γ. Using this approach, Q is determined first as a function of the A and B terms on **eq. 11**. At the same time kl_Γ should be greater than about 15 electrical degrees — I would use one of the 18- to 45-degree lines in fig. 5 if I were not insisting upon a specific gamma-rod length. Then T can be obtained from fig. 5 as a function of kl_Γ and Q. With T determined, S/D lies along the abscissa of fig. 4 where D/d = 1.0.

For example, consider a self-resonant vertical monopole antenna on a good ground plane with an assumed driving-point impedance of 34 + j17 ohms at 145.4 MHz. Therefore, R_a = 34.0 ohms and X_a = 17.0 ohms. With a polyfoam RG-8/U feedline, R_o = 50.0 ohms. Since the antenna impedance step-up ratio, H_z, must be larger than the ratio R_o/R_a (see **eq.** 14), if a value of 4.0 is chosen for H_z, in this example it will satisfy that requirement.

With H_z = 4.0, the diameter ratio D/d may be found from **fig. 3** to be 1.0. Assuming driven element and gamma rod diameters of 0.375 inch, the values for A, B and Q are as follows:

$$A = \frac{R_o X_a}{H_z R_a - R_o} = \frac{+850}{86} \cong +9.884 \text{ ohms}$$

$$B = \frac{R_o \left[(R_a)^2 + (X_a)^2\right]}{H_z R_a - R_o}$$

$$= \frac{72250}{86} \cong 840.1 \text{ ohms}^2$$

$$Q = +A + \sqrt{A^2 + B}$$

$$\cong 9.884 + \sqrt{937.78} \cong 40.51 \text{ ohms}$$

Using a selected gamma-rod length of 36.0 degrees, the length of the rod in inches is

$$l_\Gamma \cong \frac{k l_\Gamma}{0.03188 f_{MHz}}$$

$$= \frac{36}{0.03188(145.4)} = 7.766 \text{ inches}$$

When kl_Γ is 36.0 degrees, T is 0.018 ohms^{-1} (see **fig. 5**). When both T and D/d are known, the spacing-to-diameter ratio S/D can be found from **fig. 4**. In this case, S/D = 3.3. Therefore, the center-to-center spacing is

$$S = 3.3D = 3.3(0.375) \cong 1.24 \text{ inch}$$

Now, the quantities E and F may be determined, and the value of the gamma capacitor, C_Γ, calculated:

$$E = \frac{R_o}{R_a}\left[\frac{(R_a)^2 + (X_a)^2}{Q}\right]$$

$$\cong \frac{72250}{1377.3} \cong 52.46 \text{ ohms}$$

$$F = \frac{R_o}{R_a} X_a = \frac{50(+17)}{34} = +25.0 \text{ ohms}$$

$$C_\Gamma = \frac{1{,}000{,}000}{2\pi\,[E + F]\, f_{MHz}}$$

$$\cong \frac{1{,}000{,}000}{70{,}711} \cong 14.14 \text{ pF}$$

Therefore, the completed gamma-matching network consists of a 0.375-inch diameter gamma rod, about 7.75-inches long, with a 14pF gamma capacitor.

tee match

Although the previous discussion pertains to the gamma match, the same design philosophy may be applied directly to the tee match. If you want to tee-match a balanced transmission line to an isolated thin linear antenna at the driving impedance point, the driving-point impedance, Z_a', and the line characteristic resistance, R_o', are halved and the procedure for findings S/D, D/d, kl_Γ, and C_Γ for one arm of the tee-match follows that for the gamma match. The results are merely imaged to, or flipped over, to the other arm of the tee- or double-gamma configuration.

For example, to match a balanced feedline to the previously discussed homebrew array, which has an input impedance of 14 + j3 ohms, the gamma-match values on page 51 would be used in each arm of the tee-match to provide a match to *balanced* 100-ohm transmission line. If you want to use 300-ohm balanced line, use the same gamma-match design procedure. However, use R_o = 150 ohms for *each* arm of the tee.

acknowledgement

I want to thank W7ERU for his assistance. Without his computations, reasonable limits and graph plotting would have been a tedious task.

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ham radio

with the bottom wire installed 6-inches above the ground. I used poultry fence wire for this fence because each vertical wire is electrically welded to each horizontal wire, so there are no noisy electrical connections.

This wire provides approximately 0.6 square foot of surface area for each lineal foot of length. As shown in fig. 2, the total length of this wire is 190 feet. Seventy-feet of this fence is east of the tower and 120-feet is to the west (including the dog-yard fence). Additionally, 60-feet of the same fence wire is nailed to the rear of the house to augment the fence wire on the posts.

The four directions seen by the base currents provide approximately 2.4 square feet of conductor area per lineal foot of ground plane. This is reduced as the distance from the tower base is increased.

The fence wire is connected to the base of the tower with three lengths of ¼-inch copper tubing: two to the fence on the posts and one to the fence wire nailed to the side of the house. All connections, including fence wire interconnections, are made by brazing with brass rod. No buried wires or radials are used except the ¼-inch copper tubing connections which are 2-inches deep in grass sod.

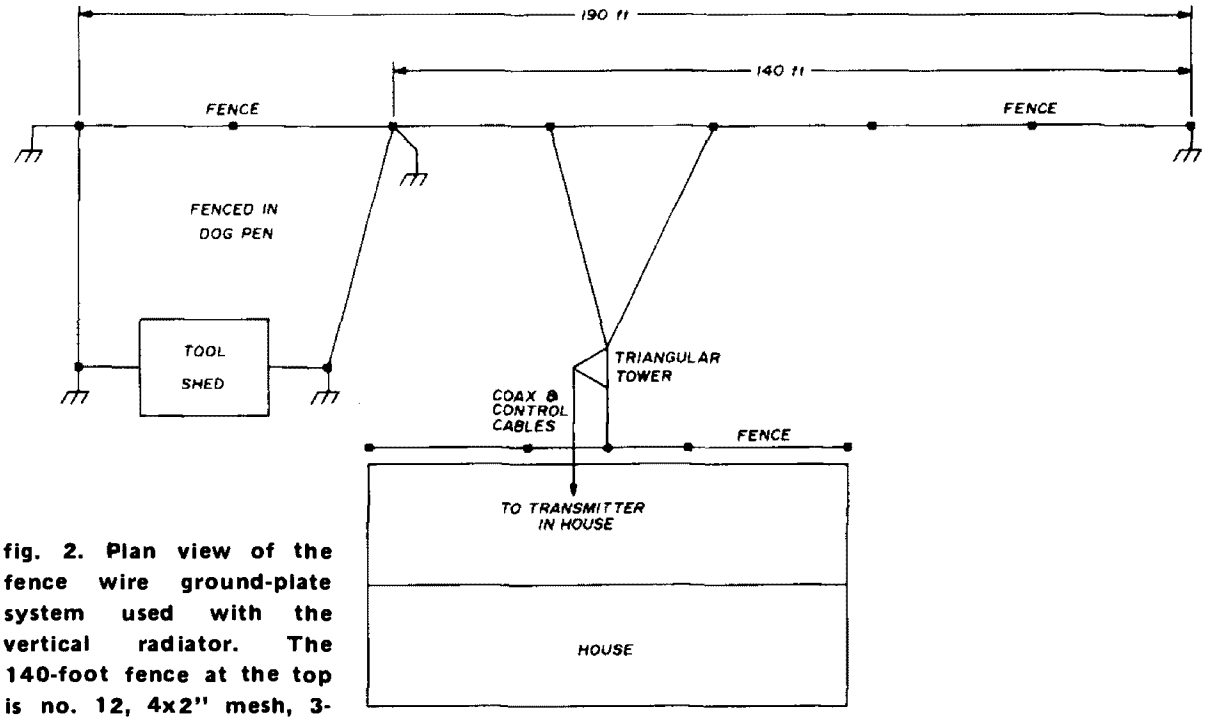


fig. 2. Plan view of the fence wire ground-plate system used with the vertical radiator. The 140-foot fence at the top is no. 12, 4x2" mesh, 3-feet high. The fence around the dog pen is no. 14, 1x2" mesh, 4-feet high. The fencing nailed on the back of the house uses the same material as in the 140-foot fence but is 60-feet long.

electrical measurements

Preliminary electrical measurements indicated that a 10 to 250-pF variable capacitor connected in series with the end of the gamma rod would cancel the inductive reactance seen at that point.

A breadboard layout was constructed and tried out at the tower base. The variable capacitor was adjusted to produce zero reactance as shown on a General Radio model 916A rf impedance bridge. Resistive readings were taken at 50-kHz increments from 3.5 to 4.0 MHz and from 7.0 to 7.3 MHz. The 80-meter band showed 5 to 6 ohms resistive and the 40-meter band showed 42 to 60 ohms resistive.

impedance matching

With the electrical characteristics available, solution of the impedance-matching problem was done by use of a graphic method (see fig. 3).[1] The method allows a solution by using only a straight edge, a

compass and graph paper. Simple arithmetic is all the mathematics that is required. Additionally, a visual choice of alternate solutions is immediately available during the graphic solution. The accuracy of constants produced is better than available equipment parts.

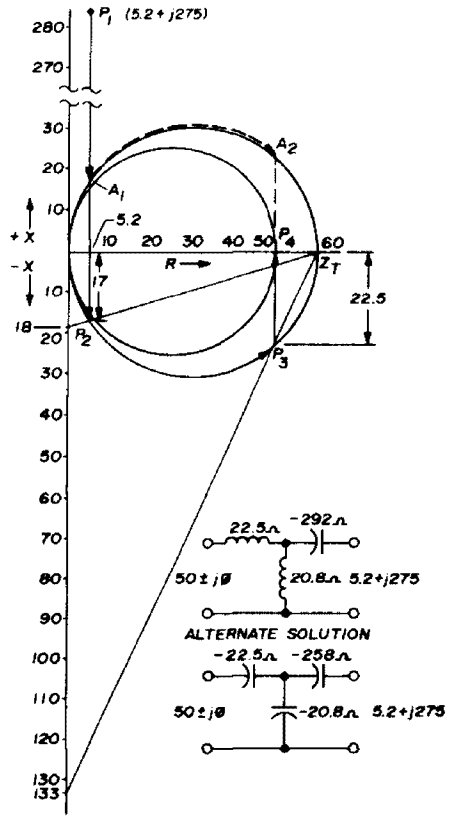


fig. 3. Graphic solution of the impedance-matching problem. See reference 1 for more information on using this technique. The input impedance of the antenna is shown on the graph at P1 (5.2 + j275). The distance from P1 to P2 is 292; since it is in a negative direction this line represents a capacitance reactance of 292 ohms. To get to the 50-ohm point, P4, you must traverse P2 to P3, representing a shunt inductive reactance of 20.8 ohms, and from P3 to P4, a series inductive reactance of 22.5 ohms. An alternate solution is to go from P1 to A1, from A1 to A2, and from A2 to P4. This provides the alternate matching network consisting of three capacitors.

The graphic solution shows that for operation on 75 and 80 meters, an inductive reactance of 20.8 ohms in shunt and an inductive reactance of 22.5 ohms in series will convert the 5.2 (average ohms) resistive component to 50 ohms for matching to the 50-ohm coaxial line. The use of ferrite toroids makes a compact low-Q L-network. By using the rf bridge while building the inductors it is possible to get them right on.

When these inductors are housed in a metal weather-resistant box at the tower base there may be some minor change in the characteristics of the gamma rod and the matching components. If low vswr is not attained re-measure the resistive component with the shield box closed. Then, make another graphic solution and rebuild the required components.

The 40-meter data shows that the resistive component is within tolerance for low vswr without need for a matching network. A dpdt rf relay switches the coaxial line from the L-network on 80 meters directly to the variable capacitor for use on 40.

observations

The tower was elevated from 22- to 32-, 40-, 48- and 54-foot levels with only small changes in the electrical characteristics. Therefore, it would appear that any tower from 25- to over 75-feet in height should be capable of being used as a grounded vertical radiator on 40 and 80 meters, with appropriate matching networks.

Most enterprising amateur antenna design enthusiasts can bum, beg, borrow or otherwise obtain the use of a radio frequency impedance bridge. Most colleges and electronics companies have such a bridge. A permanent solution to the problem of availability of an rf impedance bridge would be the construction of a Macromatcher[2] by a group or club.

The bridge used *must* differentiate between the resistive and reactive components with fair accuracy if you want to properly design the required components to couple the grounded tower, with gamma system, to the coaxial transmission line.

This vertical radiator system has relatively little loss due to ground losses and has proven to be an excellent radiator on both 40 and 80 meters. It could be used on 160 meters with reduced efficiency. In this regard, the vertical radiator would be superior to a horizontal dipole, unless the dipole was half-wavelength above the

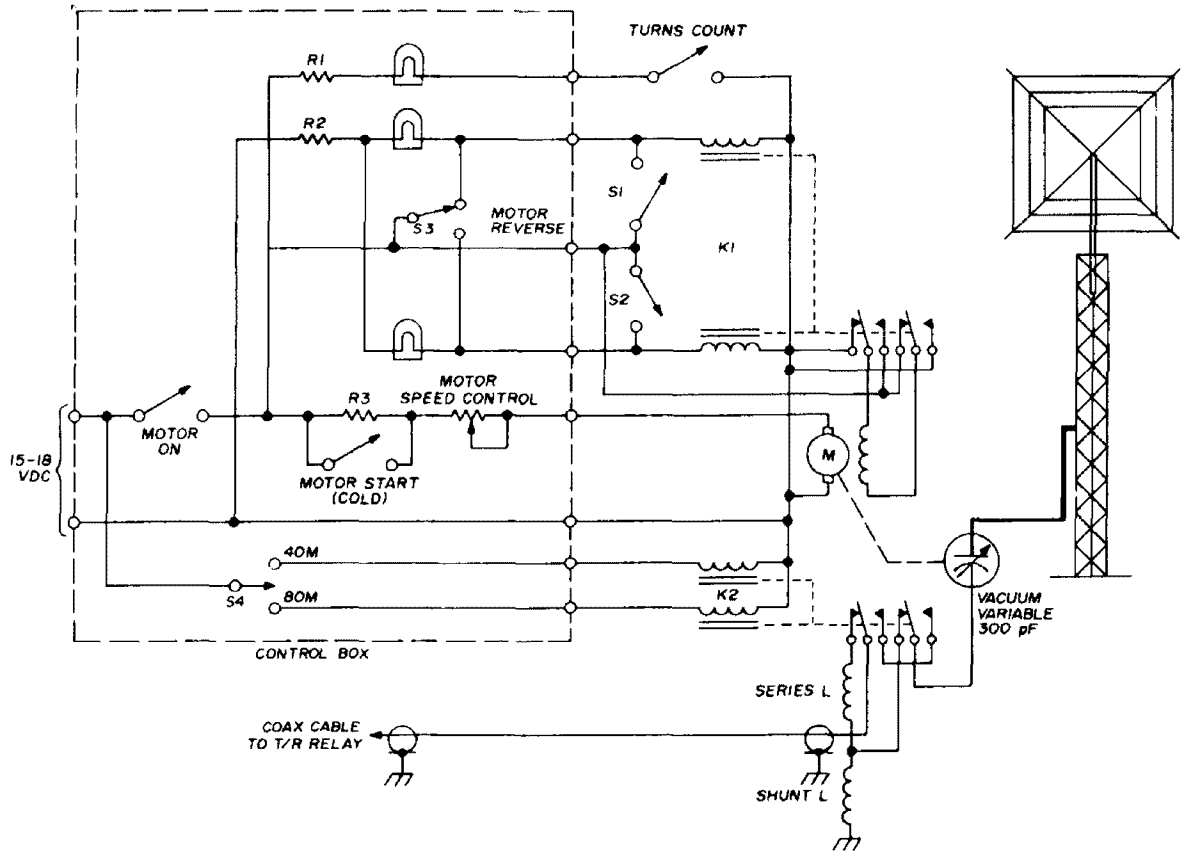


fig. 4. Alternate antenna control system which uses latching ralays to control motor direction and bandswitching. Microswitches S1 and S2 are limit switches on the vacuum variable capacitor. The turns-count switch is actuated each time the vacuum-variable shaft makes one revolution. Resistors R1 and R2 are pilot-lamp current-limiting resistors. Resistor R3 prevents current surges when the motor is cold. The motor is a 24-volt, 1/12 hp unit geared down to drive the vacuum variable. The low-Q series and shunt inductors are wound on 2" Amidon toroidal cores.

ground. Of course, an additional advantage is a lower angle of radiation on 10, 15 and 20 meters.

operation

Both forward and reflected vswr meters are used in the coaxial line. By controlling the motor driven vacuum-variable capacitor until the forward meter peaks and the reflected meter dips to minimum, the exact null-out of the reactance can be seen. The use of the reversing switch and slow speed on the geared-down motor makes it possible to get right on.

Either of two control systems may be used to provide complete control of the remote tuning system located at the base of the tower. One system was described previously[3]. A more complex control system is shown in fig. 4. This system has several features worth considering. The most important advantage is its ability to automatically reverse the motor field, hence direction of the vacuum-variable capacitor, when actuating either of the limit microswitches.

A second advantage of the control system in fig. 4 is the use of latching relays which require no holding current (and hence, no electrical noise). With these relays, once the armature has been shifted and latched by a short burst of current, it can be dropped out by another short burst of current through the unlatching solenoid.

A third remote-control system is on the drawing board and in the parts-collecting stage. This system will follow the general design of the automatic solid-state antenna tuner described by WAØAQC[4].

conclusion

The operational results obtained in the past few months prove conclusively that

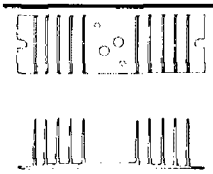
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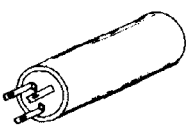
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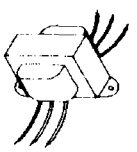


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the vertical tower antenna is an excellent omnidirectional radiator. With the control system outlined here low vswr can be obtained throughout the 40- and 80-meter bands. This broadband effect is the result of the massive tower structure and the effect of a large top-hat of quad (or Yagi) elements.

Unfortunately, there is another set of characteristics for an omnidirectional radiator when used for receiving. Because such a structure is broadband, unless high- and/or low-pass filters are provided, image frequencies and/or broadcast band harmonics could reduce the signal-to-noise to interference ratio. However, the necessary filters are also easy to design using the graphic solution technique. Additionally, signals from all directions may produce interference on the desired operating frequency.

To some, this project may seem an overwhelming task to undertake, however, for those of you who have small back yards and want to get on 40 and 80 meters without using a narrowband Z-shaped doublet or a droopy vee, may find it a necessity. Those of you who are dedicated antenna design and construction enthusiasts should find this a rewarding project.

Acknowledgement is made to Stanley Steinberg for engineering, grammatical and editorial comments, and to Stanley Dlugosz, W3EVB, for the excellent photography presented here. Last, but by no means least, apologies to an understanding wife for many delayed meals and for many piles of junk in the wrong places.

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ham radio

suitcase antenna

Complete description of a continuously-loaded multiband trap antenna that fits in a suitcase

Brian Warman, VK5BI, Cowell, South Australia 5602

Some amateurs like to take their equipment with them when they travel. Mobiling these days is very easy with modern rigs and the many fine commercial antennas which are available. It is not so pleasant though, to work mobile in today's traffic, and many traveling hams hanker for a ragchew in the comfort of their motel room at the end of the day. What is required here is a portable antenna.

A loaded whip, helical or whatever, needs a groundplane. It is also prone to detuning when moved; these considerations prompted me to look for an alternative and I decided on a dipole.

the antenna

The 5BI Suitcase is a continuously-loaded trap dipole antenna for 80, 40 and 20 meters. It is 65-inches long when assembled, and may be dismantled to fit diagonally into a suitcase. The maximum dimension when dismantled is 33 inches. I have tested it to an input power of 500 watts PEP.

Any antenna which has its radiating portion compressed is going to be inefficient. However, with today's compact high-power transceivers these inefficiencies are less important. What is required of any antenna before it can be used is resonance at the required frequency and an ability to match the transmitter output. When designing a loaded antenna it is necessary to decide on your favorite operating frequency and wind the anten-

na to suit. Any deviation from this frequency will cause a marked rise in vswr.

The design impedance is determined by the spacing between the sides of the dipole, i.e., the proximity of the two coils. In the case of the prototype the impedance was in the order of 300 ohms on 80 meters.

Before attempting to construct this antenna two pieces of test equipment are essential. Every amateur should possess an antenna bridge of some sort. I recommend the Omega Noise Bridge because of its simplicity and low cost. With this instrument it is possible to read off resonant frequency and impedance quickly and without resorting to math. There have also been a number of constructional articles on similar devices in recent amateur magazines. The other piece of test gear to have on hand for this project is a grid-dip oscillator.

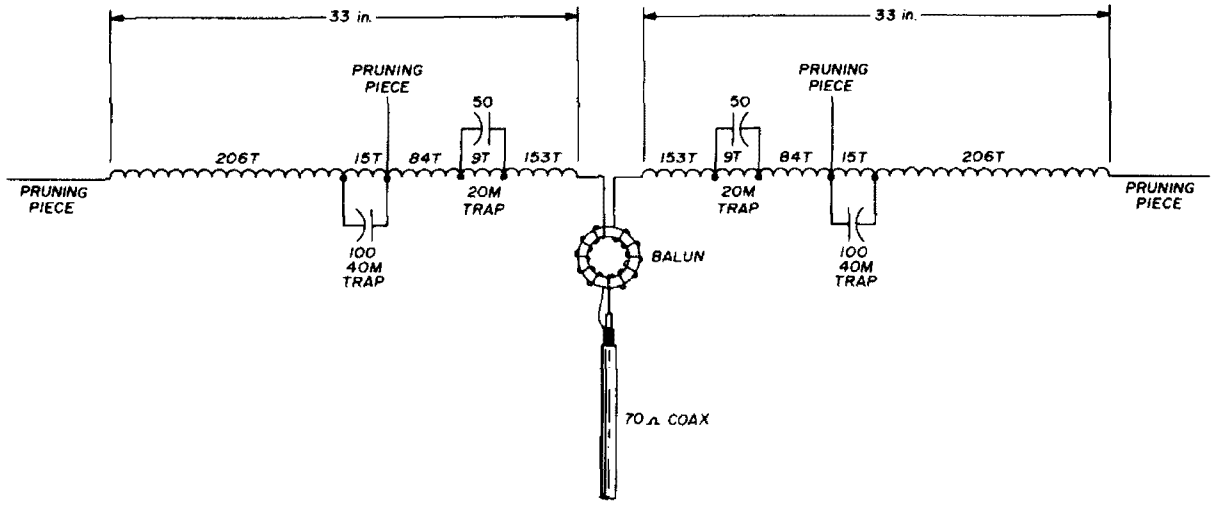


fig. 1. Continuously-loaded trap dipole is small enough, when dismantled, to fit into a suitcase. Antenna can be erected in just a few minutes.

construction

The antenna is wound on two pieces of rigid 1-inch diameter polyethylene plumbing pipe, 31-inches long. My antenna is designed to resonate on 3.60, 7.10 and 14.20 MHz, but as will be explained, it is not difficult to change the resonant frequency.

Each side of the dipole is wound identically. Starting from one end, wind 153 turns of number-24 plastic-insulated copper wire. Anchor it at 153 turns and wind a further 9 turns (for the 14-MHz trap) and anchor it again. Now, wind 84 turns for the 7-MHz section and follow this with 15 turns for the 7-MHz trap. Finally, wind 206 turns on each for the 3.5-MHz section.

The traps are resonated as follows: the 14-MHz trap is resonated with 50 pF and the 7-MHz trap is tuned with 100 pF. The capacitors I used were 500-volt micas.

The antenna is used in conjunction with a balun transformer and 70-ohm coax feedline. My balun is constructed from two pieces of Mullard FX1588 ferroxcube ring. Make two windings with number-14 copper wire, one of 10 turns and the other of 2 x 5 turns. Connect as shown in fig. 1. This provides an impedance of 50 ohms on 20 meters, 80 ohms on 40 meters and 50 ohms on 80 meters. This is well within the matching capability of most modern transceivers. Of course, if you really wanted to get fussy, a separate balun could be constructed for every band.

tuneup

It is important that, when in use and during testing, the antenna be well clear of surrounding objects. The antenna is assembled using a piece of wooden broomstick at the center, giving 2½-inch

spacing for the windings. If desired, the traps may be pruned, although I found them to be uncritical. This would be best accomplished by adjusting the trap tuning capacitance although you'll find the figures given to be fairly close.

The antenna is then connected to the bridge and the 20-meter section resonated, if necessary. This is done by connecting a short length of wire to each of the dipoles before the trap. The 40-meter section is adjusted next. In my case 15 inches were added to each 40-

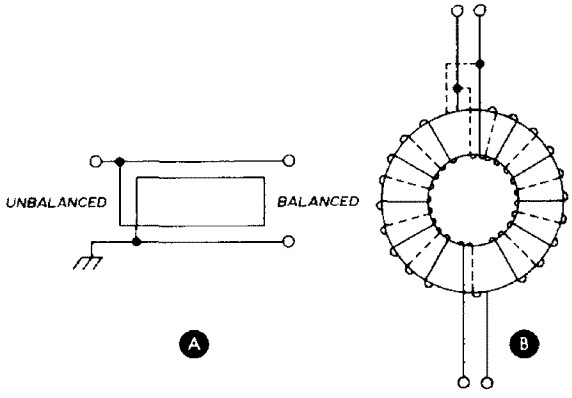


fig. 2. Circuit (A) and construction (B) of a 4:1 balun for use with the suitcase antenna. Although the author used a Mullard toroid core, an Amidon T-50-2 core with 10 or more bifilar windings of number-14 wire would be suitable.

meter section and placed at right angles to the plane of the antenna.

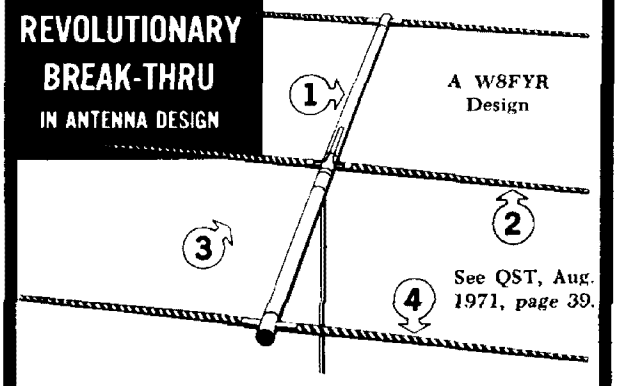
On 80 meters I finished up with 4 inches added to the end of the coil. Keep an swr bridge in the line and see that the antenna doesn't swing against any objects; this will alter the resonant frequency.

Keep people away from the ends of the dipole while you're transmitting. I haven't had any corona problems, but the antenna *talks* due, no doubt, to the electromagnetic concentration in the turns. I found the directivity to be about nil.

No difficulty was experienced with rf feedback although trouble had been expected. The transceiver I use is a Yaesu FtDX 400, which is about the same as the FtDX 560 sold in the United States.

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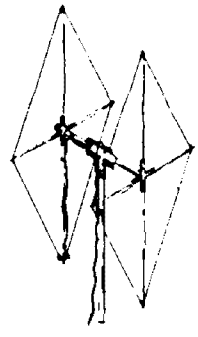
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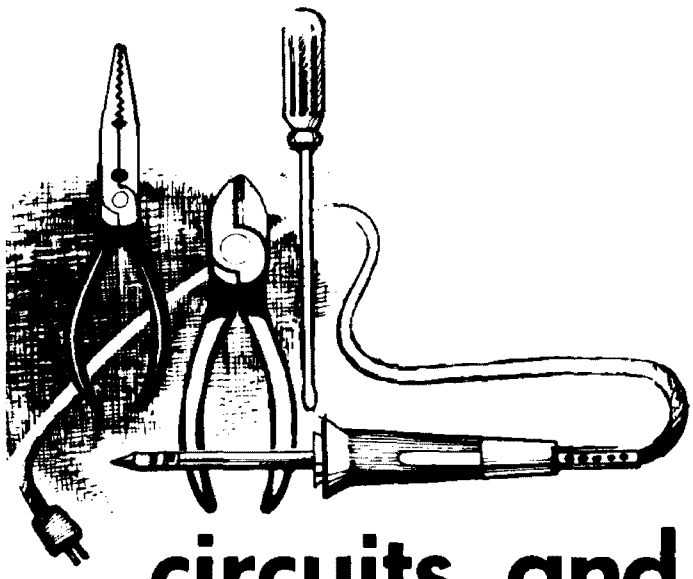
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circuits and techniques

ed noll, W3FQJ

the antenna lowdown

In the realm of ham radio antennas the second best, and even third best, will get you communicating. Unless you are telepathic you don't make any contacts by dreaming of the ultimate antenna. Too little space, too low, no money, etc., are often self- comforting excuses.

yard-high antennas

Good low antennas perform well for short and medium distance communications and can produce surprises, even in the realm of DX contacts. Low antennas can be built conveniently and economically and are less prone to deterioration by weather extremes. They perform well when they have a large aperture area and, for this reason, should preferably be full-dimension types. Also, the full-dimension type is easier to match and is less critical of design as compared to the smaller dimension DDRR types.

Take a single-element quad, lay it over horizontally, and mount it three- to four-feet above ground and you have a full-dimension low profile antenna, fig. 1.

This close-to-ground position lowers the antenna resistance of a square full-wavelength antenna to a value that provides direct match to 50- or 70-ohm line. I built versions for 10, 20, 40, 80 and 160 meters and with appropriate length trimming a direct match was obtained on each of these bands. The equation used for determining wire length was:

$$\text{Length}_{ft} = \frac{984}{f_{MHz}}$$

Resonance was found by shortening the antenna between two and three percent. For example, typical wire lengths for 40 and 80 meters were 132 and 245 feet, respectively.

Supports for the antennas were four metal fence posts. The antenna wire sizes varied between number 12 and number 16, insulated. In construction the radiating wire is passed through one hole of the antenna insulator. A short piece of

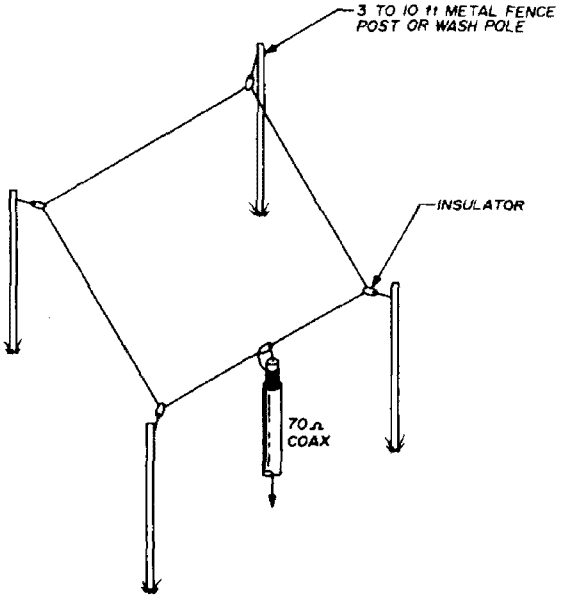


fig. 1. The low, square antenna — a horizontal quad.

wire through the opposite eye holds the antenna at each metal fence post. Typical dimensions are given in fig. 2.

On 40, 80 and 160 meters good signal reports are obtained for what is considered normal short and medium distance communications for each of the particular bands. For these ranges signal level readings compare favorably with good antennas of greater height. Performance falls off in the DX range of each of these three bands, indicating further work is needed to lower the vertical angle of radiation. As yet, I have not done any experiments with ground systems beneath the antenna, feed arrangements that include the radiating antenna and ground or tilting the low profile plane.

The economics of this arrangement is attractive because no high support structure is required. It is made from inexpensive wire, and in most cases, the length of transmission line can be made short. Low height means experimental ease. Our tests were made in the 3- to 5-feet above ground range. However, you can anticipate little change in matching if the antenna is raised up 8- to 10-feet and supported on ordinary wash poles.

Space requirements are modest and, in many situations, the erection of a full dimension antenna is possible where it is not feasible to erect a half-wavelength 80- or 160-meter dipole. For example, the 80-meter versions fit into an area that is only about 62-feet square. Performance on the 10-, 15- and 20-meter bands is poor as compared to good high antennas on these high frequencies. As mentioned previously, no experimentation has been done on ground systems that may well bring down the vertical radiation angle and improve performance on 10, 15 and 20.

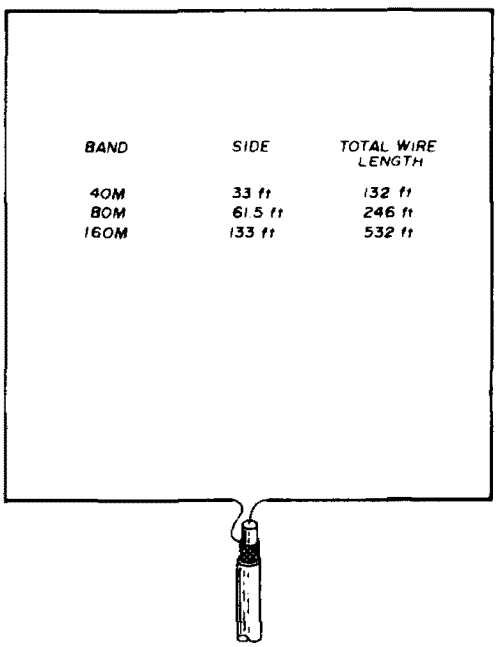


fig. 2. Dimensions for the low horizontal quad on 40, 80 and 160 meters.

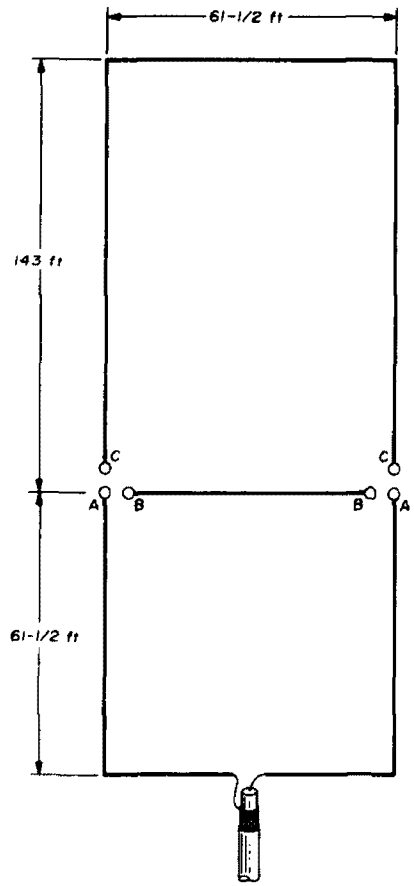


fig. 3. Combination 80- and 160-meter antenna for a long, narrow lot. For 80-meter operation, jumper A to B; for 160 meters, jumper A to C.

two-band version

The shape of such a low antenna need not necessarily be square. It can be made to fit the erection site. The directive characteristics of other than the square configuration have not been thoroughly investigated.

A two-band 80-160 combination has been checked out and performs very well on both bands. This combination consists

of a closed square for 80 meters while the 160-meter version is rectangular as shown in **fig. 3.** A good direct match is obtained on both 80 and 160 meters. Wire jumpers are used to select either 80- or 160-meter operation.

multiband open configuration

Tests were made with the *far-end open* for a square 80-meter version of this antenna, **fig. 4.** The open plan requires the use of a *tuner.* The T-network type described in the January, 1973, issue of *ham radio* permitted a match on all bands, 6 through 160 meters. Performance was acceptable on 160 meters even though the overall wire length is a bit under one-half wavelength.

If you want a direct match on 80 meters, you need only place a short across the open end of the square. If you desire full-wave performance on 160 meters and have the available space, use the same idea in setting up the antenna for all-band operation. Again, a short

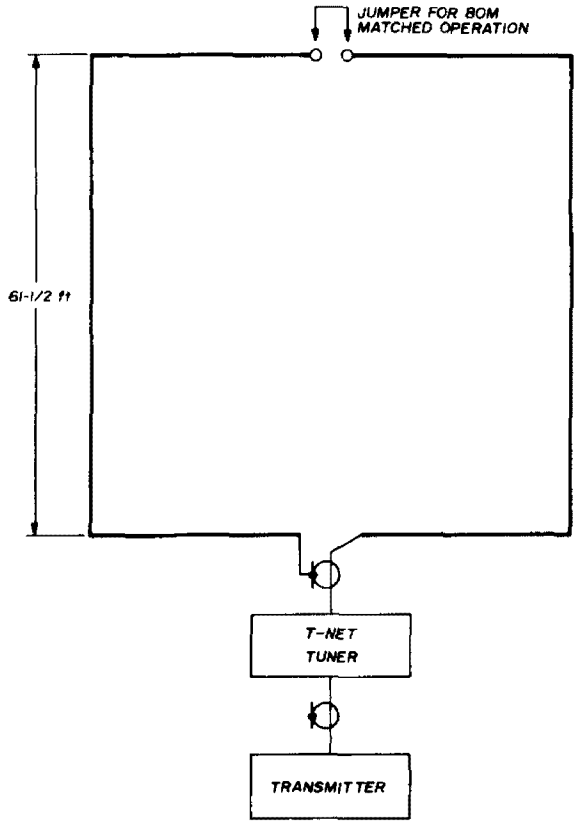


fig. 4. The 80-meter open-square antenna for all-band operation. Total length of wire in the square antenna is 246 feet, 61½ feet per side.

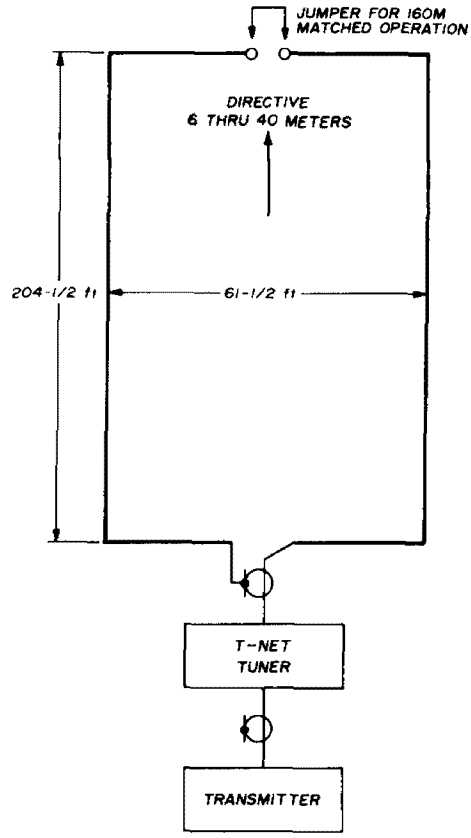


fig. 5. The 160-meter open rectangle antenna for all-band operation has some directivity on 7 MHz and higher.

connected across the opening will provide a direct match on 160 meters whenever it is desired.

The 160-meter open rectangle, **fig. 5,** also loaded on all bands using the T-network tuner. Good results were obtained on 40, 80 and 160. Due to the long sides the configuration became a good directional gain antenna on 10, 15 and 20, maximum off the far ends.

double-barrel long wire

High band results encourage further study. It is interesting that the separation between wires is about one wavelength on 15 and 20, the two bands of good directivity, **fig. 6.** This provides side cancellation.

If the far ends are terminated in Beverage fashion, how much better would the antenna perform than a regular Beverage? What would be the optimum height for a given band, pair of bands, trio of bands? What is the ratio or improvement using four wires?

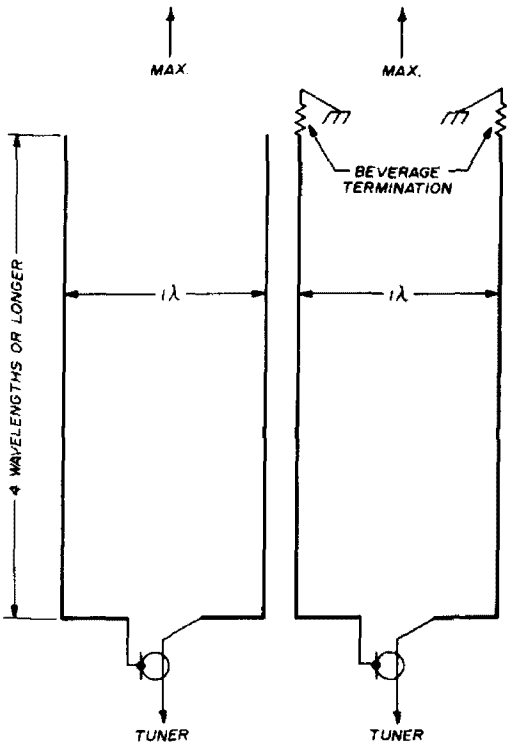


fig. 6. A new directional antenna that shows promise (see text).

There have always been exciting possibilities for low antennas. The proof is in the rather surprising performance obtained in mobile operations. Fixed installations are not so limited in size, and the performance, economics and esthetic aspects of low antennas should not be ignored.

ham radio



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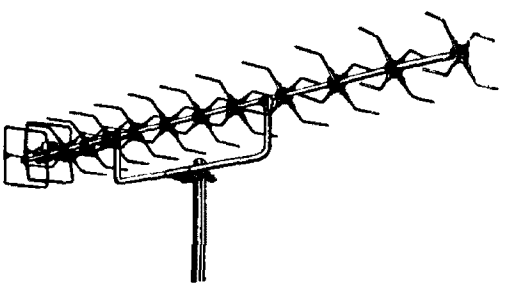
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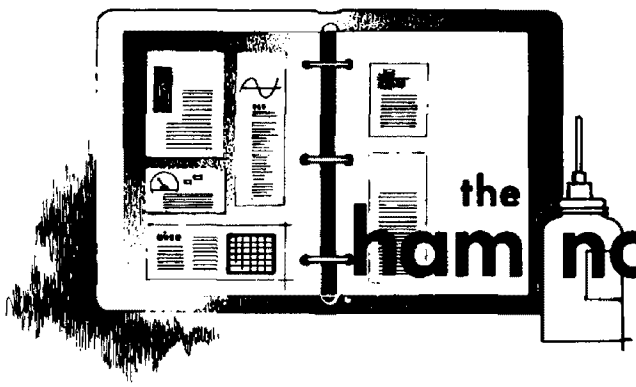
Gain — 22.1 DBi
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the ham notebook

homemade antenna insulators

With the advent of various new wire beams and antennas, there is a need for inexpensive, lightweight antenna insulators and feedline spreaders. Here is how you can fabricate these items using readily available materials.

Obtain lengths of dowel rods at your lumber supply house or hardware store. These wooden rods come in 36-inch lengths and in various sizes from less than 1/4-inch to 1-inch in diameter. For number-14 antenna wire, or smaller, half-inch dowels are adequate. For larger wire, use the 3/4- or 1-inch size. For invisible antennas and open-wire feedline spreaders the 1/4- or 3/8-inch sizes are correct.

Measure off the lengths desired, taking into account the probable useage and the size of the pan used to impregnate the rods in paraffin or bees wax. Mark the holes to be drilled (fig. 1). Various size holes are chosen for the expected wire size. When the dowel is sawed apart to make the correct lengths, the larger holes will make slots in the ends of the dowels.

Drill all holes with the smaller size drill; then re-drill with a larger size at the ends of the insulator. Saw at points X. Smooth the ends with a file or sandpaper and file or saw a slot in the ends where no hole half was made at the original dowel-rod ends.

Obtain several cakes of paraffin wax; read the instructions with care. The wax must be cautiously melted, preferably in a double boiler. If an open flame is used extreme caution must be taken to prevent splashing or other hazards that would result in a fire! An aluminum *Sara Lee* pound cake pan is just right for up to 7.5-inch long rods. The pan can be placed in another larger pan of heated water to make a double boiler.

When the wax is completely melted gently lower the drilled dowels into the pan. Under constant supervision, allow the dowels to boil and bubble for about 30 minutes. By that time the wax will be absorbed as much as possible. When all bubbling ceases, remove the dowels. They are ready for use.

The smaller hole may be used to accept the antenna wire itself, or to pass a serving wire through (in the case of feedline spreaders) to connect to either the antenna or the halyard. Larger holes

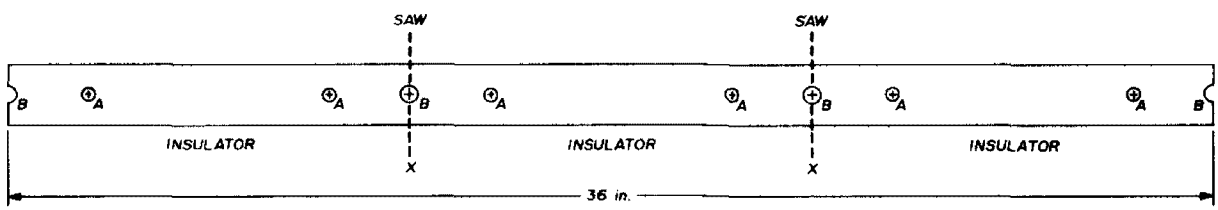


fig. 1. The 36-inch wooden dowels available at your local lumber yard may be cut into convenient lengths and boiled in paraffin or bees wax for use as antenna insulators. The holes marked A are slightly larger than the wire to be used; the holes marked B are twice as large as A.

may be used to accept a tie ring to a halyard, if desired. Keep all holes as small as practical.

A. David Middelton, W7ZC

multiband coaxial dipole

Although many amateurs use single-band coaxial dipoles, I have built a multiband unit that performs quite well on all bands, 80 through 10 meters. The layout of the antenna is shown in fig. 2. In case it looks familiar, the idea came from a similar, all wire antenna in the A.R.R.L. *Radio Amateur's Handbook.*

The coil in each of the traps consists of 21½ turns of B&W coil stock, 2-inches in diameter, 8 turns per inch (B&W 3900). The tuning capacitor is a ceramic 47-pF unit rated at 6000 working volts. The coax braid and center wire are soldered together at the ends. The braid is carefully cut at the center for the feedline — the center conductor of the feedline is connected to the braid of one antenna section while the feedline braid is connected to the braid on the other antenna section.

With the dimensions shown in **fig. 2** antenna provides very low swr over the phone segments of both 40 and 80 meters. On the higher bands the swr is less than 1.5:1 over most of the band. For best performance on the CW ends of the bands, the antenna would have to be made slightly longer. Adjust the length of the coax sections for the 7.0-MHz band, and the length of the end wires to resonate the antenna at 3.8 MHz.

H.W. Rieben, W4BDK

portable vhf ground plane

Although I am not Scotch, my resilience factor goes up and down everytime some new (and expensive) piece of ham gear is wanted, but no funds are at hand. It is my opinion that ham equipment is too specialized (and therefore limited) in today's consumer market. However, that's only an opinion and a moot point at that. I am sure the EIA would vehemently argue with me on the subject.

Since I have been in ham radio for many years — although certainly not an old timer — I really get excited when a new idea, for old equipment, comes along. Especially when it doesn't cost too much money. Take, for example, that quarterwave mobile whip standing in the corner of my room. That's the only thing left after the two-legged sharks stripped my automobile of my ham gear. Alas, it was never recovered.

Now, what could be done with a no-hole trunk-mounted whip for two meters? Well, after hunting around the house for a few weeks, an old economy type camera tripod was discovered. This tripod did not have a handle adjustment lever for the camera plane as the more common tripods do. It was constructed of brass tubing with a minimum closed length of 15-inches and a fully extended length of 48 inches. Just perfect for a three-radial portable ground-plane antenna.

The tripod mounting hole for the camera uses a Standard ¼-20 machine bolt. So, a piece of sheet aluminum (3x4x0.2") was drilled and tapped for

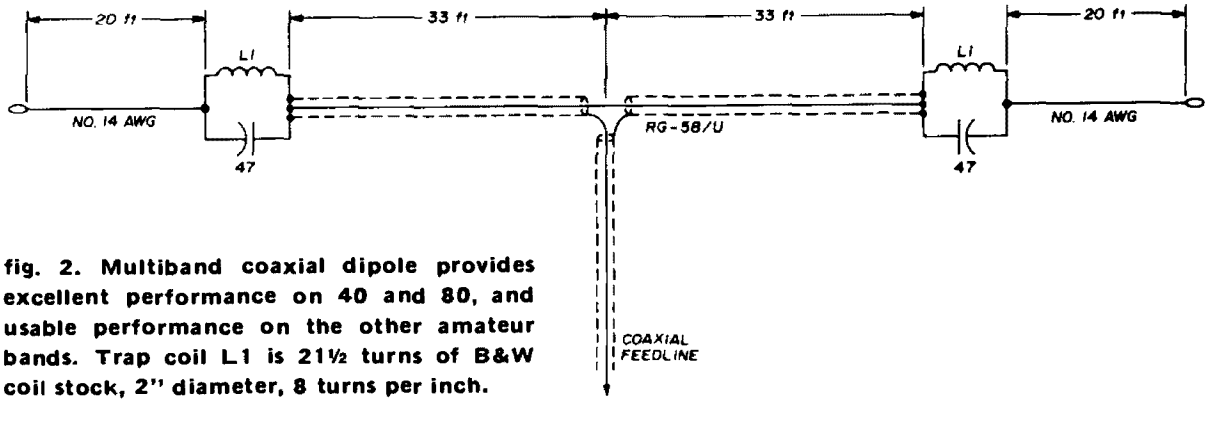


fig. 2. Multiband coaxial dipole provides excellent performance on 40 and 80, and usable performance on the other amateur bands. Trap coil L1 is 21½ turns of B&W coil stock, 2" diameter, 8 turns per inch.

¼-20 at its symmetrical center. The sheet metal was then mounted to the tripod top with a ¼-20 bolt. It resembled a miniature table with long, skinny legs. The whip could then be mounted to the table top by tightening the antenna set screws. Thus, a variable-angled-radial groundplane antenna for two meters.

From experiments using a Bird 43 watt-meter, it was determined that the lowest swr of 1.12:1 occurred when the 15-inch tripod legs were angled almost horizontally to the ground. Of course, the tripod was placed well above ground during these tests. The test frequency was 146.94 MHz.

It is feasible that this construction method would work quite well with almost any type of whip, perhaps a 5/8-wave or even CB antennas. Just vary the length of the tripod legs and the angle and see what develops.

John Sego, **K9DHD**

effective radiated power

Effective radiated power, not transmitter power, is what counts. Your ERP, especially on two meters, may only amount to less than half of your transmitter's output due to loss in the feedline. This can be reduced by installing exotic coaxial cable, costing over $100.00, or by using some alternative.

Look at the figures. Loss at 144 MHz per 100 feet of RG-58/U is 5.7 dB, RG-58/U foam, 4.1 dB, RG-8/U, 2.5 dB, RG-8/U foam, 2.2 dB, 3/8-inch Heliax, 1.3 dB, 300-ohm tv twinlead, 1.55 dB, and 300-ohm open-wire tv feedline, 0.75 dB.

Foam-filled coax has a slight advantage, but it may not be worth the extra cost, especially for lines less than 100-feet long. The solid sheath Heliax cable is much better, but a 100-foot line of this cable with the special fittings, would cost close to $100.00, so it is out for most hams.

Tv 300-ohm lead-in has low loss and since it was specifically designed for use up to 200 MHz or so, it should be best. On a cost basis it is the most for the least, but its superiority may be neglected by the average ham.

Some objections have been heard regarding its operation during wet weather, but they do not seem to be justified. There are many tv antennas operating around 200 MHz that do not seem to lose weak signals when wet, or show mismatch, although it may exist. A coat of floor wax on the twin lead will prevent moisture changes.

No difficulty in either mismatch or weak signals has been noted on two antennas I feed with this line. Open-wire 300-ohm line is the ultimate in efficiency and is unaffected by moisture. The difficulty is matching it to the usual 50-ohm input impedance termination. An impedance-matching unit is required, and if built of low-loss components it has negligible insertion loss. In fact, the ERP is usually increased due to the better impedance match.

The impedance matching unit shown in fig. 3 can match any balanced line to any unbalanced line (coax); it will even match 72-ohm coax to 52-ohm coax, if desired. This line tuner will also increase the rejection QRM generated by other services (tv, fm, etc.). It reduces harmonics by about 30 dB, and due to its perfect balance to ground, reduces noise pickup. Of course, a 300-ohm termination is required at the antenna.

This line matcher is actually easier to use than a gamma matching system. Use a 1-to-1 folded dipole, or a step-up dipole for a beam. I use 3/16-inch tubing and number-12 wire for the folded dipole on my beam. It provides a perfect match and is not at all critical.

Using the beam design shown in the *ARRL Handbook,* a beam with a gain of 6 to 8 dB is easily realized. This, with my 10 watter, provides an ERP of at least 40 watts if 100-feet of 300-ohm line is used — four times the power at less cost! If I used RG-8/U the beam would only offset the line loss and my ERP would be 10 watts. Using a half-wave dipole, I would only have an ERP of about 2 watts if

RG-8/U coax was used for the feedline.

The 2-meter line tuner consists of a 22-inch length of 1/4-inch copper pipe and a small tuning capacitor, which may be a small variable or two pieces of metal about 1½ inches square, spaced about 1/8 inch. Bend the tubing into a U about 2-inches wide and connect it as shown in fig. 3. Use a good ground at the center.

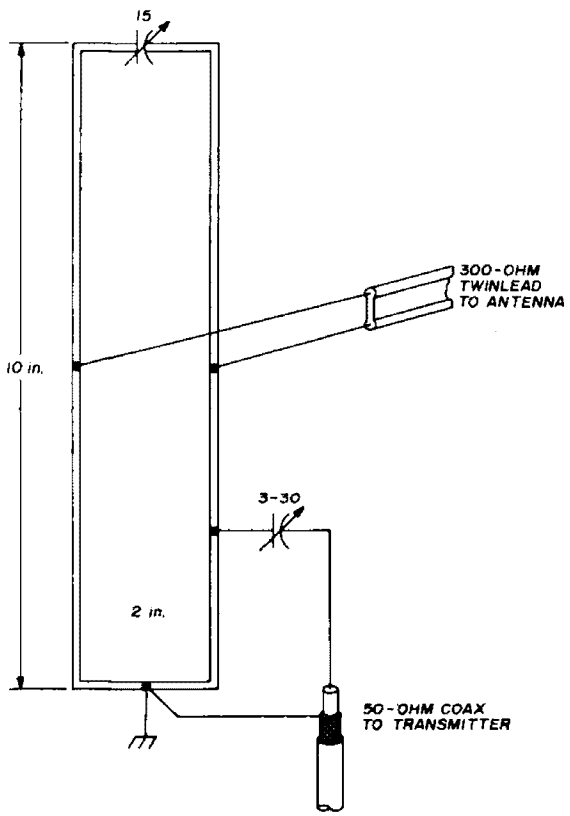


fig. 3. Two-meter line matcher provides good match between 300-ohm twinlead to the antenna and short section of coax to your transmitter. Power loss in 300-ohm plastic twinlead at 144 MHz is about 1.55 dB per 100 feet. Open wire 300-ohm line is only .75 dB per 100 feet.

Tune the line matcher for maximum forward power and least reflected with an swr bridge known to be accurate at 144 MHz.

My 300-ohm feeders are soldered on about half way up, and the 50-ohm short line to my rig is terminated with a gamma match with the tap about 5 inches up. It works — consistent daily contacts are made over 100 miles from a difficult seashore location surrounded by hills and mountains.

Alf Sheffield, VE7CB

printed circuit tool

When working with printed-circuit boards a pair of fingernail clippers makes an excellent tool for clipping off excess wire after a component is soldered in place. I have found that fingernail clippers are much better for getting into limited space than small sidecutter pliers.

Don Farrell, W2GA

code practice

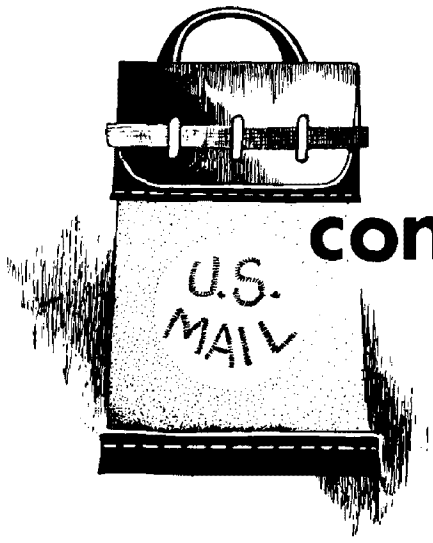
A tape recorder is a great help for improving code speed. You can record messages off the air or practice material from W1AW, then play it back at your leisure. If your machine has two speeds, for example, you may record at 3-3/4 IPS and play it back either at the same speed or at 7 IPS. This means that a recording at 10 wpm may be heard at 20 wpm if you wish.

One problem in copying code is the tendency of anticipating what is coming. You may write down letters not actually sent. This habit may be overcome by sending words like "txulg" instead of plain language. Or, plain language may be sent backwards. In this way, the listener is forced to copy only what he actually hears.

Most tape recorders don't have a reverse at normal speed, but here is how to overcome this limitation. Simply put a twist in the tape between the feed reel and the sound head. Of course, this greatly reduces the output volume so you will have to turn up the gain. It also drives the recorded channel in its *reverse* direction. Letters like R and P sound the same in both directions, but L will come out as F, and U will be D. The word "the" will be heard as "eth," and so on. The letter Z in reverse is not a letter at all. If you copy correctly, you will not write down anything because you cannot recognize it.

I have found this method valuable for increasing my code speed and strengthening the habit of copying *only* what I hear.

I. Queen, W2OUX



comments

vhf fm receiver

Dear HR:

The Allied Radio Shack stores sell a receiver listed as the *Realistic Weatheradio* cube which tunes the area of 162 MHz. By turning the converter oscillator slub approximately two turns, the radio will tune the two-meter fm band in the area of 147 MHz. I have been quite successful in doing this with little effort. Also, an fet rf amplifier would increase the usefulness considerably. Possibly a ham could build the two-meter fet transmitter described in the February, 1971, issue of *ham radio* and come up with the world's smallest hand-held fm transceiver.

Carl Markle, Jr., K8IHQ
Silver Spring, Maryland

sloping dipole antenna

Dear HR:

The excellent article by W5RUB in the December, 1972, issue of *ham radio* dealing with the DX ability of the sloping dipole antenna prompts me to write. I had been using an off-center fed 66-foot dipole, the so-called Windom antenna. Results were only so-so. It was decided to extend the overall length of the antenna to 88 feet, so 22 feet were added onto the short end, thus giving a balanced center-fed antenna. The East end was raised to a height of 75 feet. The West end is only 25 feet above ground. The antenna runs roughly East-West and is fed with 300-ohm twin line. My dc power input usually runs about 175 watts, and the measured efficiency of the final runs about 60-62 percent, putting 100 to 110 watts into the feedline.

The expected North/South maxima are quite evident; also the tilt toward the West definitely favors contacts in that general direction. Numerous VKs and ZLs have been worked, also American Samoa, KS6, and VK9, Norfolk Island. I have noticed that when abnormal (poor) conditions appear, I have had some unexpected QSOs, but only in directions North or South from my location. These odd responses to my CQs were from a Russian DXpedition to Antartica on the one hand; and over the pole from Tibet, Burma and Indonesia. These contacts took place during periods of magnetic disturbance when East-to West circuits were definitely no good. On the other hand, contacts with Europe are more difficult, S-meter readings from that area being down roughly two S units.

Summing up, the antenna performance has been surprisingly good with respect to the directions which it favors. For the economy-minded ham, perhaps two such antennas, both sloping down from the same high pole, might be the answer to effective DXing on the lower-frequency amateur bands.

Neil Johnson, W2OLU
Tappan, New York

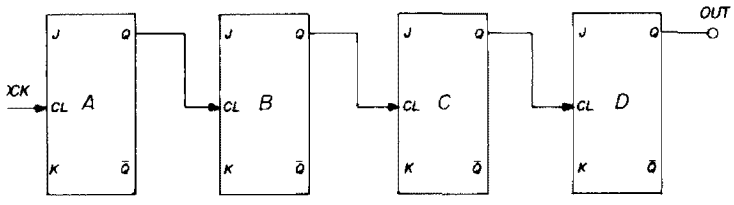
digital counters

Dear HR:

I'm writing to add a crucial remark to the information on counters by Roy Lewallen in the December, 1972, issue of

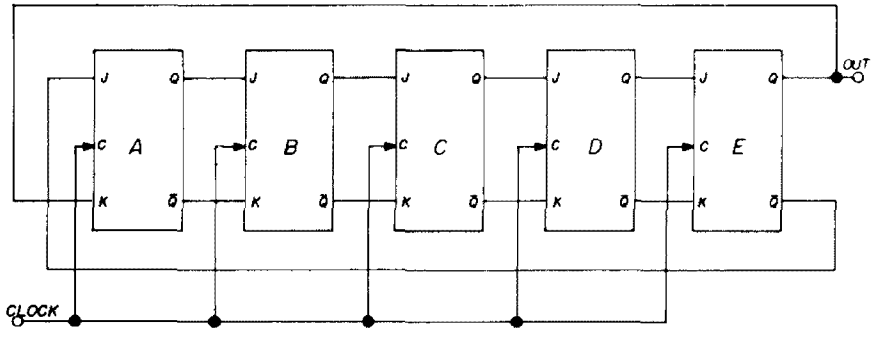
ham radio. Everything he states is correct, but his construction is missing an item that may be of interest to many potential IC users.

Consider the case of a Binary Coded Decimal counting unit made of four JK flip-flops in a configuration where the J and K inputs are both tied to +1 to cause toggling:



Admittedly, the decoding is easy enough, but consider the case of a counter going from a count of 7 to 8. Every flip-flop must change its state (A, B and C go from 1 to zero, and D goes from zero to 1). This means that the clock pulse should normally be slow enough to allow *full* propagation through the counter. In the case of the Fairchild μL923, for example, the propagation is 80 nanoseconds per flip-flop, or about 320 nanoseconds for the counter; hence, the input clock frequency is limited to about 3 MHz.

Now, consider the case where the five flip-flops are used in a "2x5" code:



The corresponding code generated is seen to be:

	0	1	2	3	4	5	6	7	8	9
A	0	1	1	1	1	1	0	0	0	0
B	0	0	1	1	1	1	1	0	0	0
C	0	0	0	1	1	1	1	1	0	0
D	0	0	0	0	1	1	1	1	1	0
E	0	0	0	0	0	1	1	1	1	1

This 2x5 code is very similar to Morse code with zero = dash, and 1 = dot. Note that between any two counts of the counter only *one* flip-flop changes state at a time. Hence, the input clock frequency is limited by only a single propagation of 80 nsec, or a limit of about 12 MHz.

I contend that this scheme is just as easily decoded as BCD, and further, in the long run is less expensive because of the slower ICs which may be purchased. I have successfully used this circuit to 10 MHz, and have tested it to 12 MHz.

Stephen R. Alpert, W1GGN
Auburn, Massachusetts

Swan cw monitor

Dear HR:

Recently, my attention was called to an apparent oversight on my part regarding the CW monitor for the Swan 350 which appeared in the *ham notebook* section of *ham radio* for June, 1972, page 63.

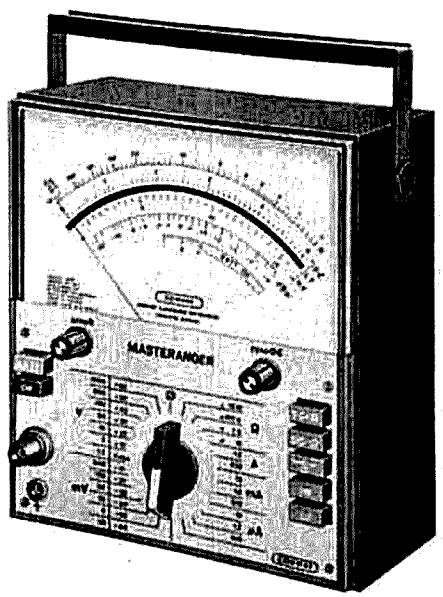
This oversight concerns the existence of R1204 (470k), R1202 (470k) and C1204 (0.1 μF) connected from pin 2 of V12 (6GK6) to point "R" in the Swan 350 audio circuitry. With these components in the circuit, V12 will be biased off during transmit and the CW monitor will not be audible. Removal of the connection between pin 2 of V12 and R1202 (or complete physical removal of all of these components, if desired) will eliminate this problem. No ill effects should be noted by this change.

I removed these components years ago during the trial installation of one of the factory CW monitor modifications and *never* re-installed them since no detrimental effects were noted. All other comments received from builders of the monitor have been favorable, especially those concerning the absence of chirp or yoop during keying.

Paul K. Pagel, K1KXA
Enfield, Connecticut

new products

high-impedance multimeter



The new solid-state Conway Masteranger multimeter provides a comprehensive range of measurements not previously available on a portable instrument of this type.

The high input sensitivity of the Conway Masteranger permits accurate measurement of very low-level voltage and currents. Input resistance on most voltage ranges is 100 megohms, voltage sensitivity is 1.5 millivolts full-scale, ac of dc, and the current range of the instrument is 0.15 μA to 1.5 amperes. Accuracy is 1.5% of full scale. The Masteranger will also measure rf voltage up to 1000 MHz.

In addition to voltage and current measurements, the Conway Masteranger measures resistance to 10,000 megohms, and has 13 decibel ranges from -80 to +66 dBm. The built-in overload protection circuitry protects the instrument to ±1200 volts on all voltage ranges; current and resistance ranges are similarly protected.

The Masteranger can be used as a null detector with high sensitivity of 20 μV per division (or 1 nanoamp per division), thus enabling use as a null detector for any bridge application or for fm discriminator alignment. When used with a calibrated microphone, the sensitive solid-state multimeter can be used as a sound-level meter. And, at very low current measuring ranges, the unit is capable of measuring contact potentials between dissimilar metals. These are just a few applications where the high sensitivity pays off.

The instrument may be powered from the self-contained battery pack or power-line operated from 115 or 230 volts ac, 50/400 Hz with an accessory power supply. Other available accessories include a 50 kV high-voltage probe, 1000-MHz rf probe, 150-ampere current shunt, capacitive high frequency voltage divider (300 volts maximum to 1000 Hz), coaxial T-connector for measuring vswr to 1000 MHz and a peak-to-peak measuring probe allowing factual quantitative p-p measurement.

The Conway Masteranger is supplied in a high-impact cabinet plus leather carrying case and batteries, input coaxial cable, two banana plugs and clips and an instruction manual. The instrument is priced at $150.00. For more information, write to Conway Electronic Enterprises Ltd., 88 Arrow Road, Weston, Ontario, Canada, or use *check-off* on page 126.

More Details? CHECK–OFF Page 126

radio handbook, new edition

Seventy years ago farsighted experimenters were communicating by wireless using spark transmitters and magnetic detectors. Today the science of radio communications has been revolutionized by solid-state devices and integrated circuits.

The completely new and up-to-date 19th Edition of the *Radio Handbook* was written especially by William Orr, W6SAI, to keep the amateur radio enthusiast informed of the latest principles and equipment encompassing the broad radio communications field. In fact, this handbook is recognized as the leading independent authority in the field of radio amateur high-frequency and vhf communication, covering more than three decades of development in the art of electronic communication.

The book contains authoritative, detailed instructions for designing, building, and operating all types of radio communications equipment. A complete understanding of the theory and construction of all modern circuitry, semiconductors, antennas, power supplies, full data on workshop practice, test equipment, radio math and calculations is privided.

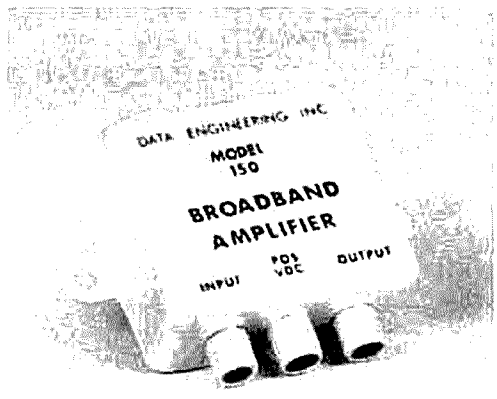
In addition, the coverage includes construction information on new high-frequency linear amplifiers of 1- and 2-kW PEP output, a solid-state LED-readout receiver, a high-performance two-meter moonbounce converter, solid-state vhf fm amplifiers, etc.

There are 976 information-packed pages supported by numerous diagrams and photographs. A glossary of terms is provided to identify the symbols used throughout the text. This invaluable reference is a must for both novice and advanced radio amateurs, electronics engineers and technicians – literally anyone interested in the popular *radio* communications field.

Topics covered include introduction to radio; direct-current circuits; alternating-current circuits; semiconductor devices; vacuum-tube principles; vacuum-tube amplifiers; radio-frequency power amplifiers; special circuitry for vacuum tubes and semiconductor devices; single-sideband transmission and reception; communication receiver fundamentals; generation and amplification of radio-frequency energy; rf feedback; frequency modulation and repeaters; radioteletype and specialized transmission and reception; amplitude modulation and audio processing; radio interference (RFI); equipment design; station assembly and transmitter control; mobile and portable equipment; receivers, converters and transceivers; exciters and transceivers; high-frequency and vhf power amplifiers; power supplies; radiation, propagation and transmission lines; antennas and antenna matching: high-frequency directive antennas; vhf and uhf antennas; high-frequency rotary-beam antennas; electronic test equipment; the oscilloscope; construction practices; radio mathematics and calculations.

Hardbound, 976 pages, $14.95 ($17.95 in Canada) from Comtec Books, Greenville, New Hampshire 03048.

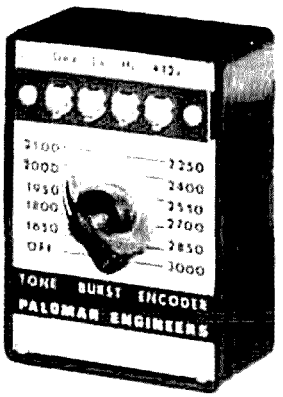
broadband preamplifier



Data Engineering's new Model 150 broadband preamplifier provides the user with the opportunity to improve the sensitivity of his hf receiver for a-m, ssb, CW or fm. Covering a frequency range of 1 to 30 MHz, the unit features 36-dB gain at 1 MHz, dropping to 19-dB gain at 30 MHz. The unit features a maximum 3-dB noise figure.

More Details? CHECK–OFF Page 126

TONE BURST ENCODER



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3 CHIP
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7430	.35
7440	.35
7441	1.30
7442	1.00
7446	1.50
7448	1.25
7450	.35
7451	.35
7453	.35
7454	.35
7460	.35
7472	.40
7473	.55
7474	.40
7476	.55
7480	.50
7483	1.15
7486	.65
7489	1.00
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741(MINI DIP)	.45
747	1.10
709	.30
710	.40
711	.40
723	1.00

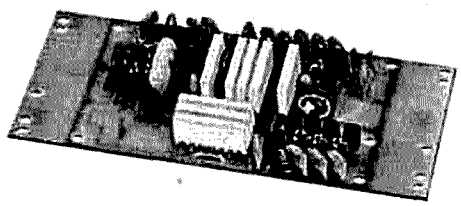
CMOS

C04001	$ .75
C04002	.75
C04011	.75
C04012	.75
C04023	.75

The unit is powered by 12 Vdc and its small size lends itself to installation within the cabinets of existing receivers. The new preamp sells for $17.95. An adaptor for 150- to 250-Vdc operation is $2.95.

Additional information is available from Data Engineering, Inc., Ravensworth Industrial Park, 5554 Port Royal Road, Springfield, Virginia 22151 or by using *check-off* on page 126.

motrac tone kit



The new Alpha sub-audible tone encoder/decoder has been especially designed for use in the Motorola Motrac series of two-way radios. The heretofore expensive and difficult task of converting so called "non PL" Motracs to the use of tone is made simple and practical with the Alpha unit.

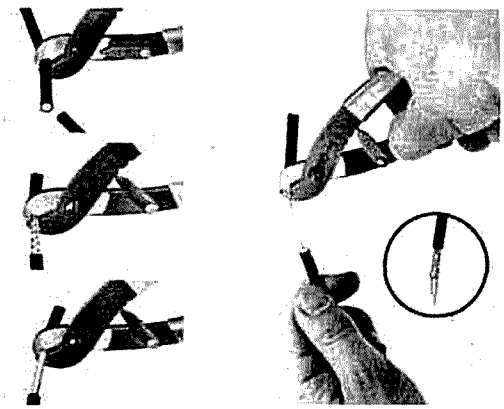
The new unit utilizes thick-film hybrid modules that contain all the active circuitry used for the encoding and decoding of tone. These thick-film modules fit into a unique gold-plated edge connector on the printed-circuit board allowing direct plug-in-plug-out operation for ease of maintenance, modification or change of tone frequency. The thick-film hybrid technique also makes possible an exceptional degree of reliability under severe environmental extremes including high vibration and temperature. The frequency determining modules are laser trimmed to the precise frequency required and are therefore not subject to the reliability problems of reeds and the frequency stability problems of tunable types of tone.

The Alpha SS-80J/192 is completely compatible with all Motorola, General Electric and RCA sub-audible tone systems and is available in standard or special frequencies from 20 to 250 Hz.

More Details? CHECK–OFF Page 126

Provision has been made to accommodate up to six tone frequencies which may be electronically switched if required. The unit also has the capability of automatic revert for common encode or common decode configuration. The multi-frequency circuitry can be provided from the factory or added in the field. Complete step-by-step instructions are provided. For additional information write to Alpha Electronic Services, Inc., 8431 Monroe Avenue, Stanton, California 90680, or use *check-off* on page 126.

coaxial cable stripper



Xcelite has just added to its line of quality professional hand tools a coax stripper/cutter, designed specifically for use with the popular RG-59/U coaxial cable. Featured is a three-position selector lever. With the lever in position 1, the hardened and ground blades cut cleanly through the jacket, shielding and dielectric without fraying to expose the undamaged conductor. Position 2 removes the jacket and shielding, while position 3 strips the jacket off without damage to the shielding, or dielectric. With the selector lever disengaged, the entire coax cable can be cut neatly to length. Handles have cushion grips for user comfort and to lessen fatigue.

The no. 590 Coax Stripper/Cutter has a list price of $4.75. It is available nationwide through Xcelite's local distributors. Additional information may be obtained by writing for Product Bulletin 572L, available from Xcelite Incorporated, Orchard Park, New York 14127, or use *check-off* on page 126.

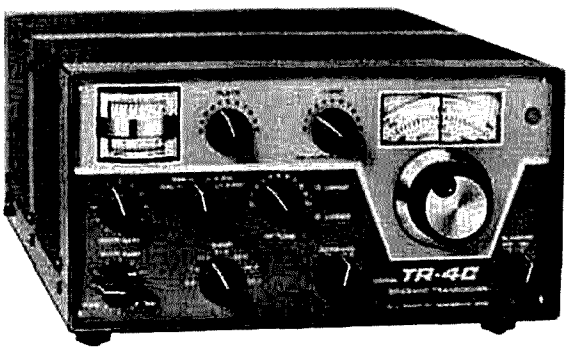
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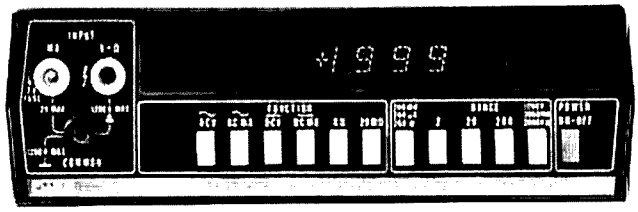
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More Details? CHECK–OFF Page 126



New digital V.O.M. works well in near field environment. Only $299.

Here's the best low cost digital voltmeter ever made for broadcast and communication use. It's got all the resistance range, voltage resolution, high ac accuracy you'll ever need plus 30 second warm-up to full accuracy. Fluke's new Model 8000A measures in 26 ranges ac/dc volts, amps and resistance from 100 μV to 1200 V, 0.1μA to 2 A, and 100 milliΩ to 20 megΩ. Basic dc accuracy, 0.1%. Full year guarantee. Option choice includes rechargeable battery pack, printer output, deluxe test leads, HV probe, RF probe, 600-amp ac current probe, carrying case, dust cover and rack mount. Unique self-zero eliminates offset uncertainty. Electronics are securely mounted in high-impact case. Service centers throughout U.S., Canada, Europe and Far East for 48-hour turnaround repair.

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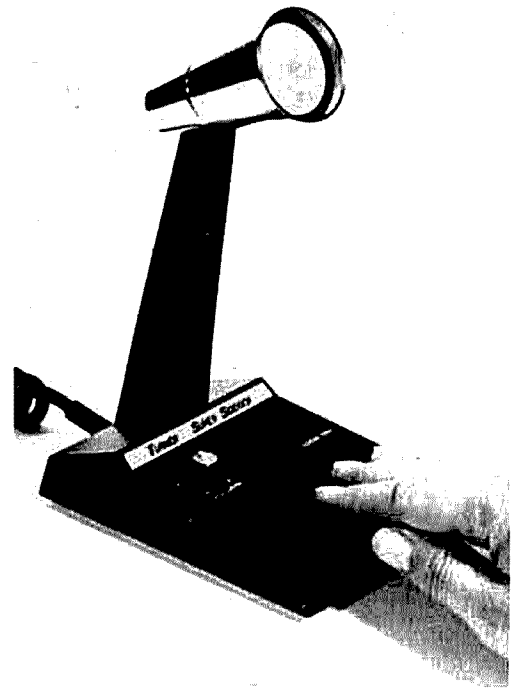
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	List	Our Price ppd.
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4CX250B	$ 35.60	$ 17.00
4X150A	$ 23.15	$ 11.00
2BP1	$ 24.00	$ 12.00
829B	$ 20.00	$ 10.00
8252/4PR60B	$115.00	$ 37.00
8187/4PR65A	$ 50.00	$ 16.00

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communications microphone

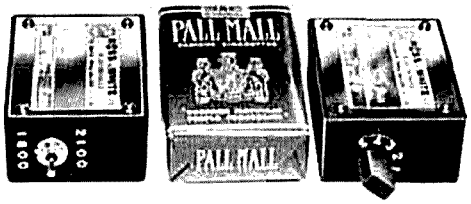


Turner's new amplified *Super Sidekick*, a base station communications microphone, has two gain controls: one on the top for normal volume adjustments and another on the bottom for matching the Turner *Super Sidekick* to various transceivers. This microphone has been designed specifically for the amateur, business band and communications field. It uses a dynamic interior for extreme ruggedness and durability and is unaffected by temperature and humidity. It has a built-in IC amplifier which provides a perfect impedance match with all a-m/ssb transceivers.

The amplifier gain control on the bottom of the Turner *Super Sidekick* uses a rugged die cast case. The base and neck are black with a polished bright chrome head. The cable is a three-conductor coiled cord. The microphone is activated by pressing down on the touch bar and can be locked on by moving the slide lock forward. The Turner *Super Sidekick* has a list price of $80 and is available from the Turner Division of Conrac Corporation, 909 17th Street N.E., Cedar Rapids, Iowa 52402. For more information use *check-off* on page 126.

More Details? CHECK—OFF Page 126

tone encoders



Powered directly by any 12-Vdc 2-meter fm transceiver, the new Ross and White tone-burst encoders allow automatic entry into tone access repeaters. Keyed by closing the transmitter microphone switch from the 12-Vdc keyed source, the encoder generates a half second tone burst which modulates the transceiver and activates the repeater.

Two models are available. The two tone model TE-2 sells for $39.95, and the five tone model TE-5 sells for $49.95 postage paid. Both are sold on a 10-day trial, money-back guarantee basis. Installation is simple with the complete instructions provided for your make and model of transceiver. The battery powered models will continue to be available for hams preferring that arrangement.

For full data including specifications, write direct to the Ross and White Company, 50 West Dundee Road, Wheeling, Illinois 60090 and ask about tone burst encoders. You can get the same information by using *check-off* on page 126.

solid-state general-coverage receiver



A new and improved version of its popular S-120 series a-m and short-wave table radios has been introduced by Hallicrafters. The new model S-125, called *Star-Quest II*, provides the user with an all-transistor, completely self-contained

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MODEL ACA-1 AUDIO COMPRESSOR features 45 DB compression range ■ Flat 20-20,000 Hz response ■ Extremely low distortion ■ Front panel compression meter and in/out switch ■ Accepts both high and low-impedance mikes ■ Easily installed in mike line ■ 110-volt a.c. or 12-volt d.c. operation ■ Only 5″ W x 2¼″ H x 4½″ D.

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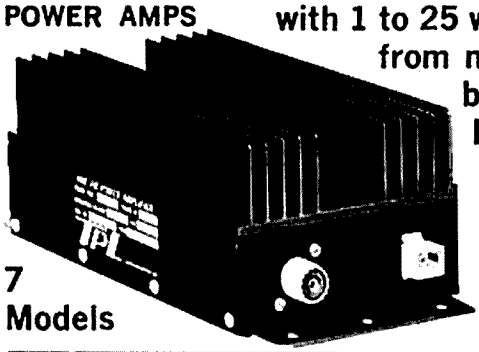
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a-m broadcast and general-coverage short-wave receiver.

Features include agc, bfo, illuminated slide-rule dial, logging scale, 1-watt audio output, bandspread, built-in speaker and provision for direct operation from 12 Vdc for portable use. The UL approved unit comes with an external antenna and offers continuous coverage from 550 kHz to 30 MHz in four bands.

The *Star-Quest II* is designed for the novice, beginner swl, hobbyist and world traveler. With built-in 117 Vac supply it is priced at $59.95. More information is available by using *check-off* on page 126 or by writing to The Hallicrafters Company, 600 Hicks Road, Rolling Meadows, Illinois 60008.

dry desoldering tool

Solder removal is fast, economical, and convenient with the new "Soder-Wick," dry desoldering tool. Used in conjunction with an ordinary soldering iron, it quickly removes solder from all sizes of electronic joints and connectors.

Used in the computer, aerospace, telephone and communications industries, "Soder-Wick" is useful for initial building, repairing, or rewiring of circuits or for salvaging parts from bargain computer boards or surplus chassis.

No special equipment is needed to remove solder from integrated circuits, printed circuits or telephone connections. Simply touch "Soder-Wick" to the heated joint and solder is drawn up immediately. Flux contamination is eliminated, and residue if any, is non-corrosive and non-conductive. In just one second, solder can be removed for as little as half a cent per connection.

Rolls of "Soder-Wick" fit conveniently into tool box or pocket, and are available in four sizes to match the joint and quantity of solder to be removed. Each roll contains five feet of wick and sells for $1.49 per roll.

For further information, contact Jensen Tools and Alloys, 4117 North 44th Street, Phoenix, Arizona 85018, or use *check-off* on page 126.

More Details? CHECK–OFF Page 126

75 cents

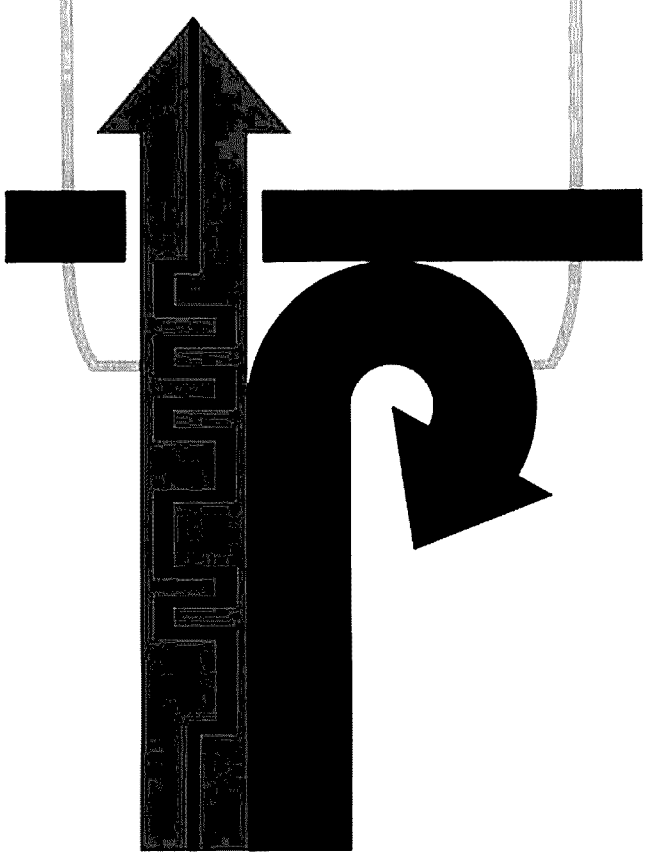
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on
communications
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ham radio magazine

JUNE 1973

digital RTTY autostart



this month

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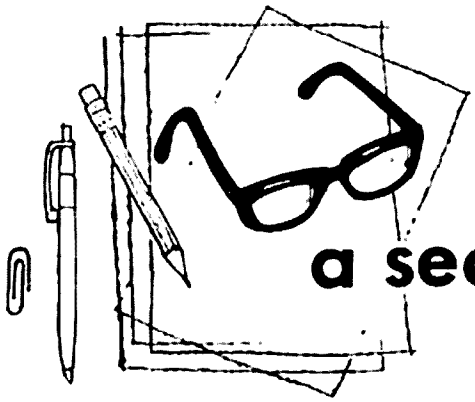
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a second look

by jim fisk

In July the United States Postal Service will issue a quartet of postage stamps saluting Progress in Electronics. The four stamps, which were announced at an IEEE luncheon in New York in late March, span electronics development from Marconi's turn-of-the-century spark coil to the transistor. The four multi-colored stamps, in denominations of 6-, 8-, 11- and 15-cents, will be issued in New York City on July 10th.

The 8-cent stamp of the series commemorates the transistor and shows transistors and other components mounted on a printed-circuit board. As we all know, the transistor revolutionized electronics, provided the basis for integrated circuits, and opened up the door to many new products, including modern digital computers, communications satellites and manned space probes.

The 6-cent stamp, intended for post cards, shows Guglielmo Marconi's famous spark coil and spark gap. Marconi, who often described himself as an "amateur," demonstrated in 1895 that by grounding the transmitter he could increase the distance over which an electrical wave could be transmitted. He accomplished this by greatly increasing the power of Hertz' spark-gap transmitter and devising a high, earth-connected antenna. One terminal of the spark gap was connected to ground while the other was connected to the antenna, thus increasing the amount of radiated energy.

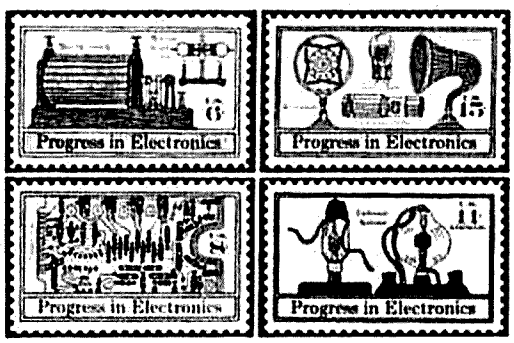
With this transmitting system and a Branley coherer, he was able to send wireless messages over a distance of 1-1/2 miles in 1895. In 1901 he succeeded in sending a Morse signal, the letter S, from Poldhu, England to St. Johns, Newfoundland, and in 1903 he sent a complete telegraph message across the Atlantic. By 1907, the inventive Italian's findings had enabled a trans-Atlantic wireless service to be established between America and England, and in 1909 his success was crowned when he was awarded the Nobel Prize for physics. It is fitting that amateurs will be able to use the 6-cent Marconi stamp for sending their QSL cards within the United States.

The 11-cent airmail stamp shows Dr. Lee DeForest's Audion vacuum tube. In 1907 DeForest introduced the control grid to Fleming's two-electrode oscillation valve, making it possible to build electronic amplifiers and vacuum-tube transmitters. In 1915 Dr. DeForest used his Audion to transmit the human voice from Arlington, Virginia to Paris. Later, he developed the talking motion picture, facsimile and television.

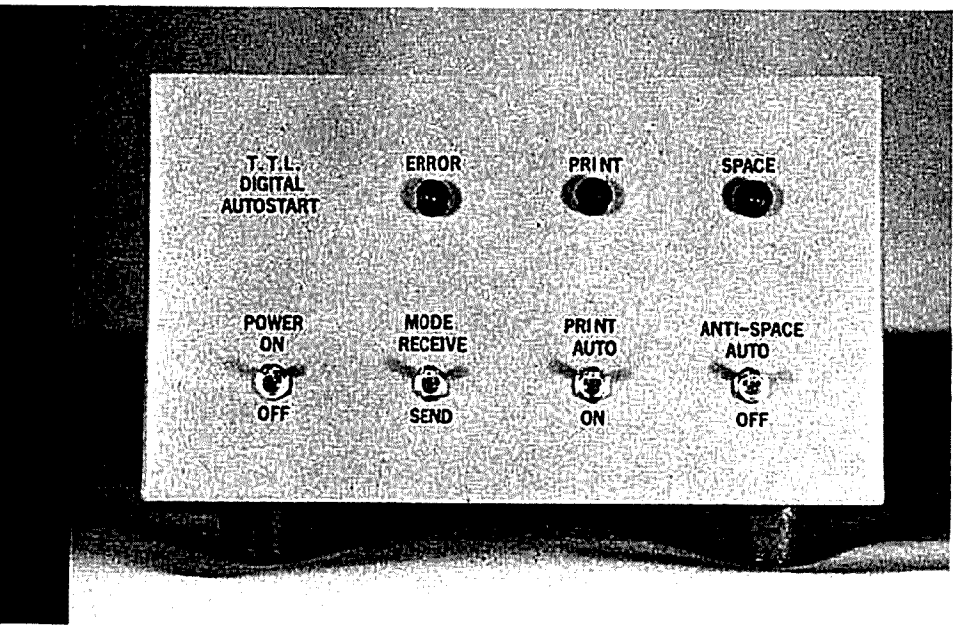
The 15-cent stamp for international surface mail features three nostalgic electronic elements — an early microphone, a gooseneck speaker and a vacuum tube — combined with a tv camera tube.

Radio amateurs desiring first-day cancellations may make requests to "Electronic Stamps, Postmaster, New York, New York 10001," enclosing proper remittance, with the request postmarked no later than July 10th. Price for the set of four stamps serviced is 40 cents. A specific stamp or stamps may be ordered for first-day cancellation.

Jim Fisk, W1DTY
editor



Progress in Electronics



digital RTTY autostart

An automatic printer control that can be used with popular demodulators without disturbing existing circuits

Bert Kelley, K4EEU, 2307 South Clark Avenue, Tampa, Florida

Autostart is an automatic printer control system that turns on the printer to copy teletype signals and turns it off when the transmission ends. An ideal system should respond only to teletype signals and ignore noise, CW, or carriers on or near the operating frequency. This not only permits trouble-free automatic station monitoring but also makes manual operation easier.

Most autostart systems now in use[1,2] are carrier operated with one exception.[3] The equipment responds to any carrier at or near the mark frequency as well as to a signal divided in time between mark and space frequencies (RTTY signal). Some circuits have a refinement called anti-space, which measures the time a signal occupies the space frequency. If this time is substantially longer than an RTTY character, the printer is placed in mark-hold. This feature offers some protection against CW stations on the space frequency and prevents the printer from running open at any time. Although these demodulators may be complex, they still depend on the selectivity of the receiver, additional audio filters, and the duty cycle of

the received signal for what discrimination they have against unwanted signals. Carriers, slow CW, and sometimes noise will turn on the printer.

autostart features

The new digital autostart system described here has the following features:

1. Positive discrimination is obtained between teletype and any other mode. The circuit checks the start and stop pulses (previously done by K5ANS).[3] However, it also checks each of the five RTTY bits that comprise the letter encoding. Since each letter is checked seven times, it is more difficult for a non-RTTY signal to escape detection by the digital logic.

2. The unit features easy, one-connection design adaptable to demodulators such as the popular ST-3, -4, -5[4] or -6 without disturbing existing circuits.

3. Inexpensive popular TTL integrated circuits are used. Circuit boards are available.*

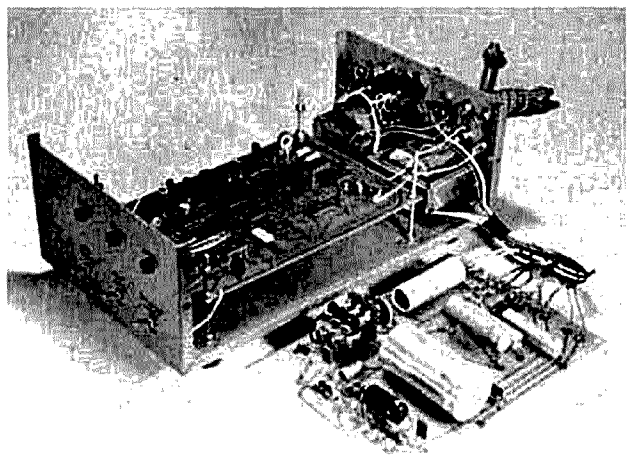
4. The digital logic, motor-control circuits, power supplies, and teletype loop circuits are contained in one package. This simplifies construction by eliminating the need for adapting to any existing power supply, loop, or motor-control circuits.

Some of the TTL logic circuits used in the autostart are shown in fig. 1. One package, U1, is Motorola RTL; the rest are Texas Instruments or equivalent 7400N series TTL logic.

operating principles

Teletype letters are sent as a series of timed pulses that have on and off conditions similar to digital logic pulses. A start pulse, which is always a space, is followed by five coded pulses or bits, which make up the letter. These bits are followed by a stop pulse, which places the printer in mark-hold ready for the next letter. This form of transmission is ideally suited to examination by digital methods.

The start pulse is a 22-ms spacing condition at 60 wpm. This pulse turns on a timing circuit, or clock, which generates



View of disassembled unit.

pulses in synchronism with the incoming signal. Each start and stop pulse is verified for polarity by the logic, and the five information bits are placed in storage and examined a moment later to determine if any change has occurred between the stored bit and the incoming signal. This decoding takes place in the center of each 22-ms interval and no change should occur. If a change occurs, an error pulse is generated. Four TTL NAND gates detect errors from (a) false starts due to noise or CW, and (b) defective stop pulses due to CW. The two storage error detectors detect changes in timing or polarity in the incoming signal. These errors are added and trigger a pulse stretcher connected so that it discharges a capacitor each time it fires.

Meanwhile, the same capacitor is charged by pulses from a similar circuit

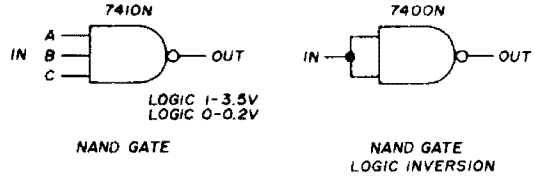
*A set of two epoxy, plated, undrilled circuit boards are available from K4EEU for $12.00 postpaid in the USA. A complete parts kit, less boards, may be obtained from Truman Boerkoel, K8JUG, Stotts-Friedman Company, 108 North Jefferson, Dayton, Ohio 45402. Write for free parts and price list.

triggered by pulses that pass examination by both the false start (FS) detector and the character gate. A circuit monitors the voltage level across this storage capacitor and turns on the printer when the voltage exceeds approximately 1.9 volts. With noise input to the teletype demodulator, errors spilling out of the four gates keep the capacitor charged at about 1 volt. When a signal appears, the capacitor will charge high enough to turn on the printer within the time required for eight characters to pass. Usually no useful print is lost due to this turn-on requirement, since most operators open each transmission with several surplus characters such as blanks, line feeds, spaces, or duplication of call letters. At machine speed, the printer will turn on in less than 1.3 seconds — probably faster than most operators could reach for a manual start switch. Shutdown is equally positive: about three characters are printed in rapid succession before printer lockup.

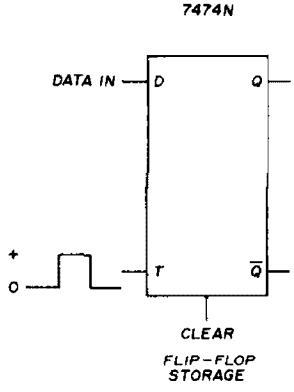
design considerations

Emphasis was placed on RTTY versus other mode discrimination rather than print evaluation. The printer will turn on whenever the logic recognizes an RTTY signal. This action may be through, or spaced between, interference. Under these conditions garbled copy may be printed, as if the printer were operated manually. Print evaluation was considered impractical, and the low threshold was chosen because garbled signals will often contain enough accurate print so that the

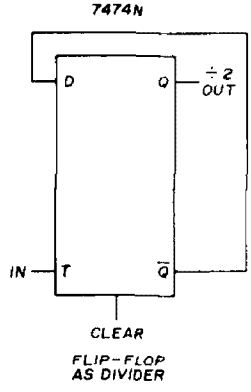
fig. 1. Logic functions of some of the TTL circuits used in the digital autostart.



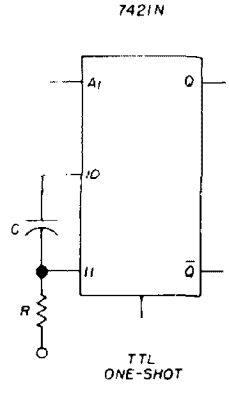
A, and B and C must be at logic 1 to get logic 0 output. For any other combination output stays at logic 1



Type D flip-flop used as storage device. When cleared, Q is at logic 1 and Q is at logic 0. Status of data at D input is transferred to Q output at positive-going edge of toggle pulse and is held in storage regardless of further changes in date input until cleared, or at next positive-going edge of new toggle pulse



Type D-flip-flop used as divider. By connecting $\overline{Q}$ to D input, device may be operated as a divider



A_1 input normally is positive and triggers the multivibrator when driven to ground. The positive output pulse appears at Q output and is inverted at $\overline{Q}$. Length of pulse increases with value of capacitor and/or resistor. The IC has an internal 2k timing resistor at pin 9, which may be connected to 5V or another resistor between pin 11 and the 5 V supply

substance of a message can be understood.

A dividing line must exist between print and nonprint, or what constitutes a recognizable RTTY signal. This grey area is particularly troublesome when propagation conditions are marginal; fading conditions make ordinary autostarts cycle on and off. Therefore a print hysteresis circuit was developed, so that evaluation standards are changed once the system goes into print. The printer will continue

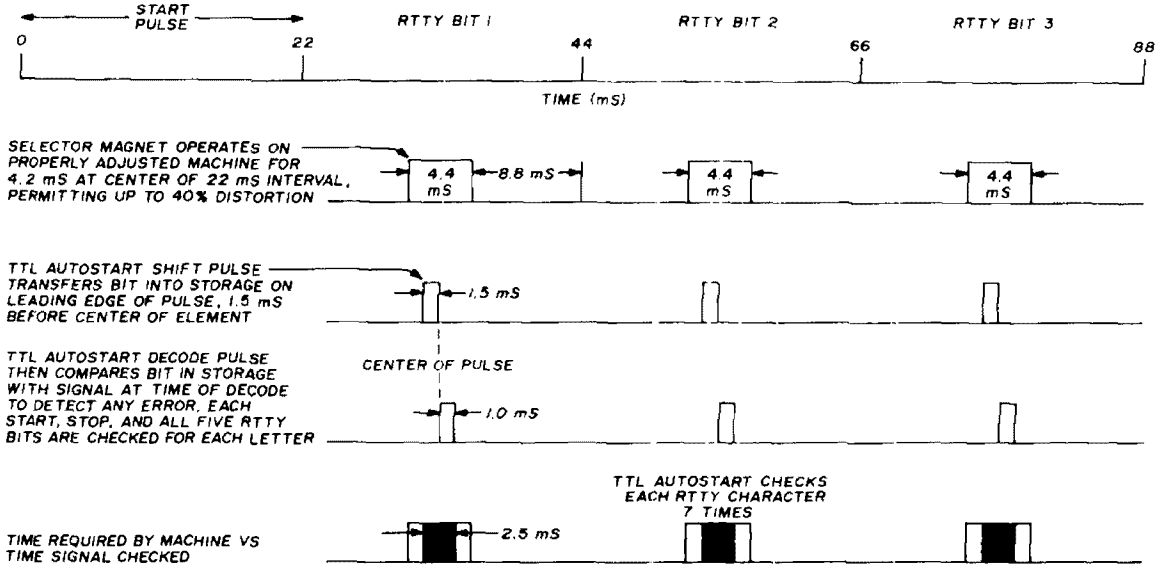


fig. 2. Timing relationship of pulses showing sequence of operation.

amounts of distortion — up to 40 percent in a well-adjusted printer[5] (see fig. 2). The print selectors operate during the middle 4.4 ms of each 22-ms interval at a point established by rangefinder adjustment. Print evaluation should take place at this time. The TTL autostart evaluates the signal *inside* the 4.4-ms interval (actually for only 2.5 ms).

storage circuit

The memory-capacitor storage circuit is the result of considerable experimentation. Several circuits were tried with digital-counter storage. These circuits have two operational deficiencies for this application. First, they have a finite capacity. When this capacity is exceeded, these counters start over at zero count rather than when a high amount of errors have accumulated that should be in storage. Second, there is no relationship with time. A carrier on the mark frequency can lock up the count indefinitely and keep the teletype motor running, yet the transmitting station may have stopped typing long ago. No doubt there are more complex methods of solving this problem.

One of my principal objectives was to keep the autostart as simple as possible consistent with required performance. A simple memory capacitor error storage circuit proved to be the solution. It will

to copy a signal that may deteriorate from the initial condition. Full anti-space protection is retained as well as positive printer control at the end of transmission. The printer motor does not cycle on and off unless copy is lost completely. This, of course, reduces loss of copy due to the eight character turn-on requirement.

distortion

An autostart system should not reject signals the printer is capable of copying. If it does, it adds errors of its own to the copy. In a practical system, the received signal is subject to varying amounts of distortion during transmission and reception due to a variety of causes — keyboard adjustment, relays, the demodulator, filters, and signal conditions, to name a few. Fortunately, the machine is mechanically designed to handle large

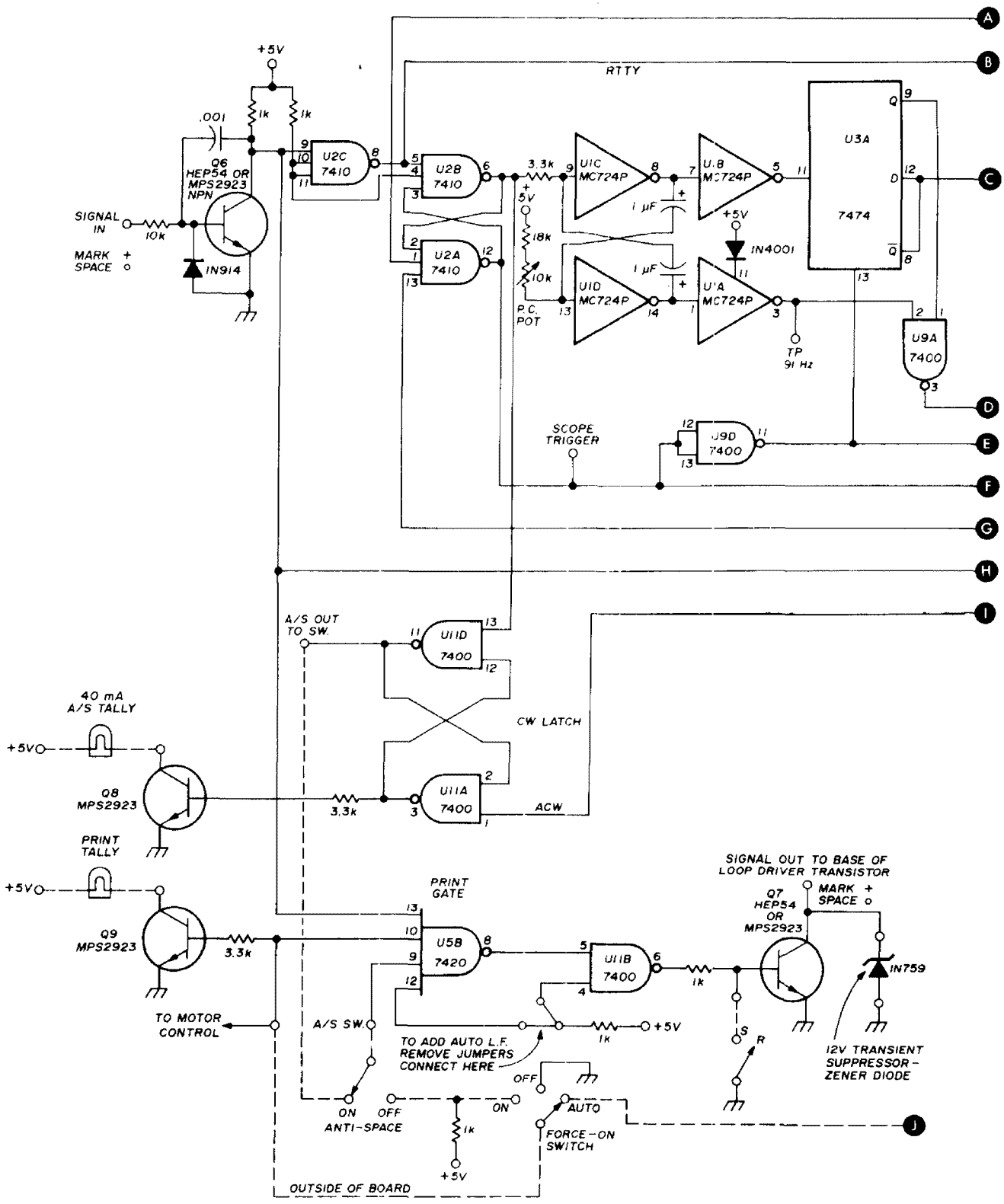
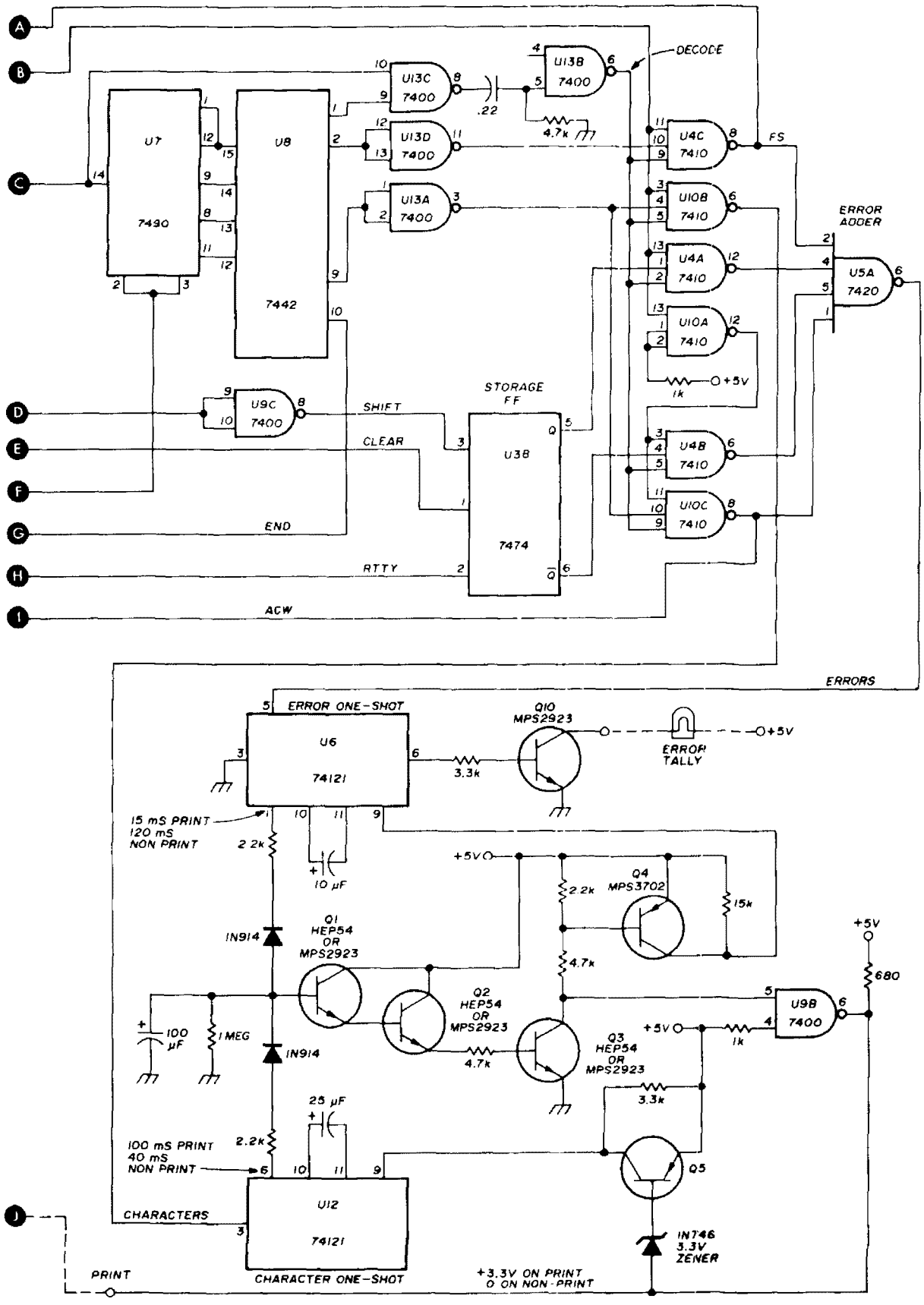


fig. 3. Logic diagram of autostart circuit board with input and output transistors.

not overflow, uses a minimum of components, and has the bonus feature of an idle line turn-off. If no typing occurs for about one minute the capacitor discharges and the printer goes off. This feature accommodates the slowest typists but turns off the printer when stations send only call letters then start tuning up on frequency.

operation

During standby, in mark hold, a positive voltage will appear at the input jack (**fig. 3**). The start pulse starts the clock



oscillating at 91 Hz. U8 then generates pulses that coincide with the start and stop pulse (see fig. 4). Seven shift pulses are generated in U9, which shifts each incoming pulse into storage. The decode pulse generator, U13, generates seven 1.0-ms positive-going pulses positioned near the center of each 22-ms interval.

Five TTL NAND gates examine the signal. Since they operate in a similar

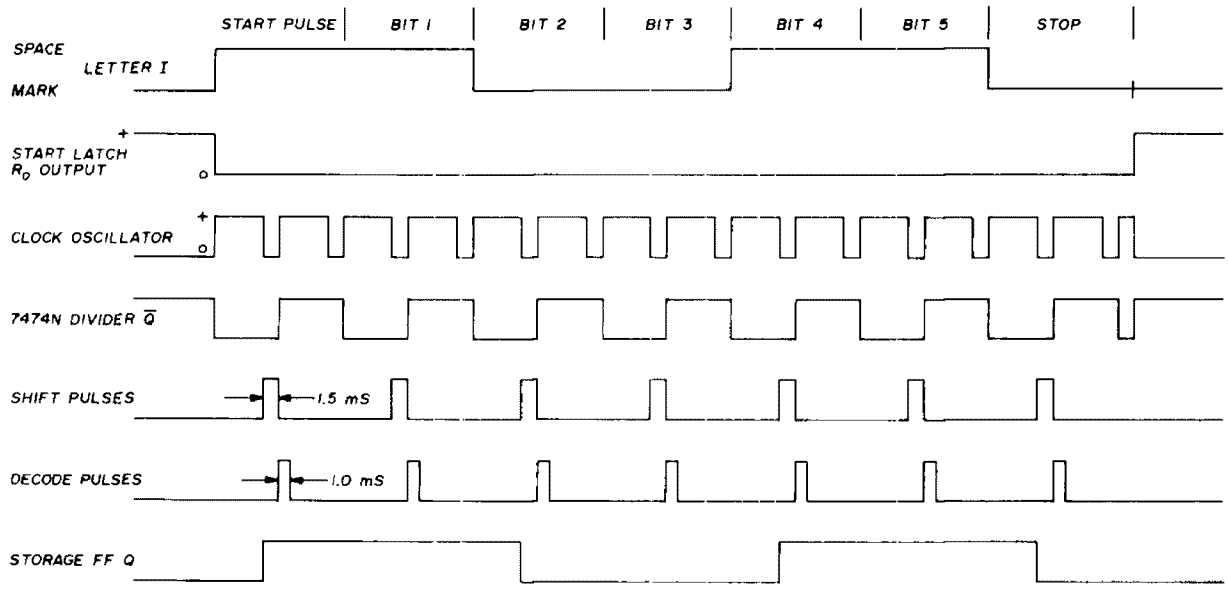


fig. 4. Waveforms of logic circuits during formation of the letter I.

manner, only one gate will be explained. One section of the three-section gate, U4, checks the start pulse. Normally the voltage at U4-8 will remain positive and will not change unless all inputs to this gate are positive at the same time.

At the start pulse the reset/set (R/S) latch, U2, sets; the clock starts; and its pulses advance in synchronism with the incoming signal, so that U8-2 is low at the same time the incoming signal is low at U4-11. U13 is connected as an inverter, so U4-10 becomes high. At the center of the 22-ms interval the decode pulse from U13-6 becomes positive. If, at this time, the incoming signal is valid RTTY, the start pulse will be low at the base of Q6, high at the collector, and low at U2-8, which is connected to U4-11. As long as U2-8 stays low, there will be no output from U4. But if the clock had been turned on by a noise pulse, the polarity of the input signal would probably be different at this time; and, since all three inputs to U4 are positive, a negative-going pulse would appear at U4-8, resetting latch U2-1 and stopping the clock.

This pulse at U4-8 is too short to see on most oscilloscopes since it lasts only long enough to reset the fast TTL logic. It may be lengthened, for verification purposes, by temporarily tacking a 1-μF capacitor between U4-8 and board ground.

When the decode pulse ends, the false start (FS) gate is disabled for the rest of the character. The other four gates require a positive decode pulse and two other positive voltages for any change in the output. Since the gate outputs are all positive, the output of U5-6 is normally low.

The incoming teletype signal is also applied to U3-2, the storage flip-flop. It is placed in storage by the positive-going edge of the shift pulse from U9-8, and the

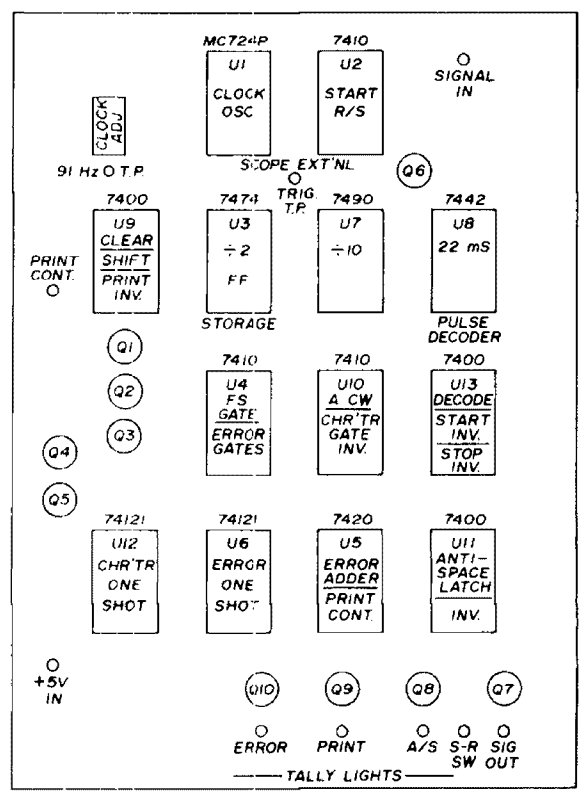


fig. 6. Logic board package layout.

signal status appears at U3-5, the Q output. This stored signal is used for comparison with the input signal to detect any changes.

The incoming teletype signal is also connected to U5-13, the print gate, which gates the signal to the following printer magnet circuit, or places the printer in mark-hold. If we assume the logic has passed an incoming signal as valid and places the system into print, then U9-6 will be high, Q9 will turn on, and U5-10 will be positive. If the anti-space latch is off, U5-9 will be positive along with U5-10, so any polarity changes in the input signal appear inverted at U5-8. U11-6 inverts the signal again to drive Q7. Two unused inputs to U5 and U11 may be used to add an automatic line-feed generator.[6] The CW latch, U11, is instantly operated by any spacing condition during the stop pulse, turning on Q8, and inhibiting print by placing a low on U5-9.

The gated and controlled signal leaves the logic board at Q7 collector and enters a second board. Q7's collector load resistor is on this second board as well as the power supplies, motor control, and loop circuitry (fig. 5). This loop is similar to that of the ST-6 and has the advantages of magnet overdrive, freedom from relays, and positive-negative FSK. The motor-control circuit is operated from the same voltage that turns on U5-10, the print gate. The diode and small capacitor at Q11 base provide a motor turn-off delay to keep the motor running for a short time between transmissions.

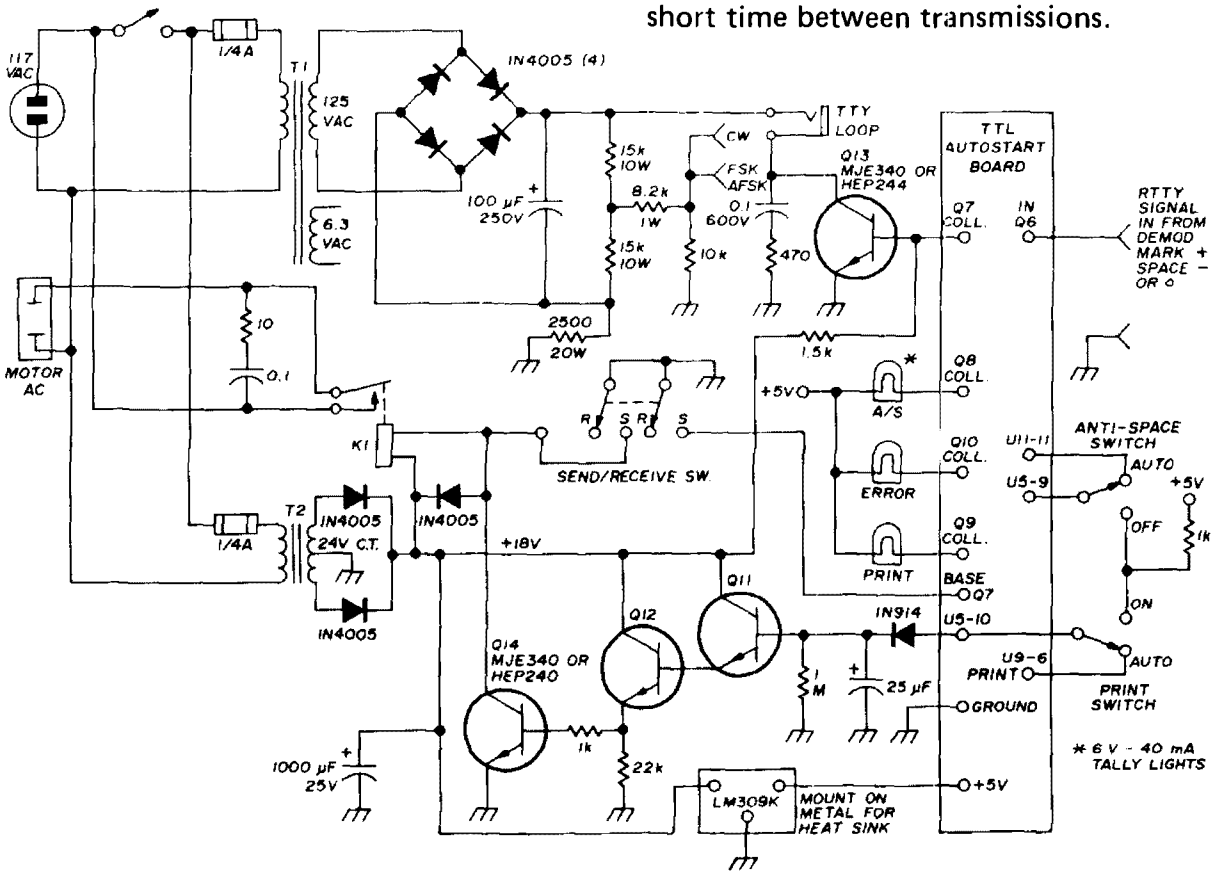
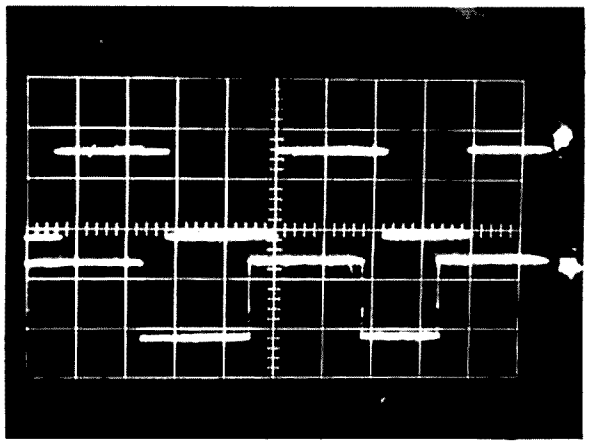
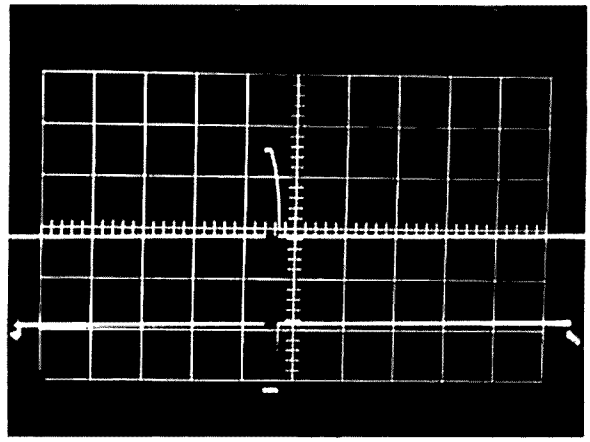


fig. 5. Schematic diagram of power supply and motor control circuitry. Relay K1 is a Potter and Brumfield dpdt KA11DG, 12-volt coil, or Olsen SW-557. Q11 and Q12 are HEP54, 2N2222, MPS2423 or MPS2924. Transformer T1 is an Allied 6K28HF. T2 is a Stancor P8180.

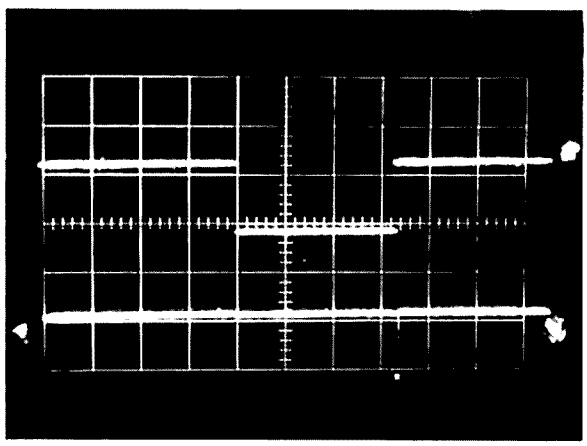
The complete unit is enclosed in a 4 x 7 x 12-inch Minibox with input and output connectors, printer jacks, and an LM309K regulator mounted on the rear apron. A sketch of the autostart board package layout is shown in **fig. 6.** The two circuit boards are stacked on long machine screws, which were obtained from toggle bolts. The power supply is mounted on top because of heat considerations. Extra wire should be provided as a service loop so this board can be removed and operated while probing voltages and waveforms in the logic. This will elimi-



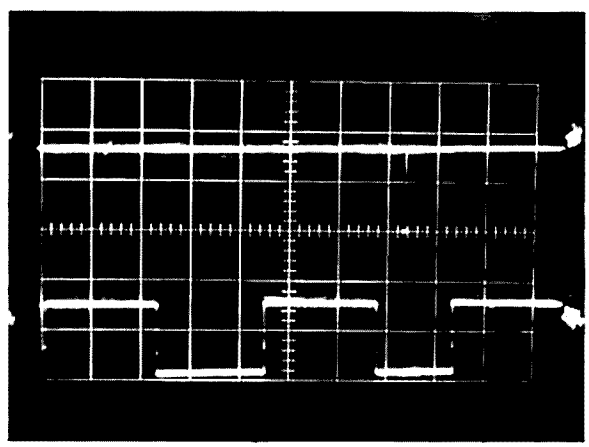
Incoming RTTY signal (bottom trace) is shifted into storage flip-flop and appears delayed and regenerated at U3-5 (top trace). 1.5 millisecond after being transferred into storage, top waveform is compared with bottom waveform (which may be static, CW, or noise and not necessarily the RTTY signal for the letter I as pictured). If a change in polarity occurs, an error pulse is generated.



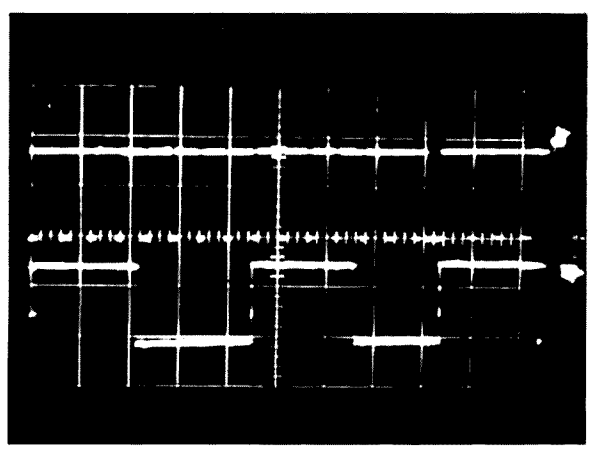
Top trace is a decode pulse; bottom is a pulse at character-gate output U10-6. Decode pulse sets timing. For this pulse to be present, denoting a valid character, a positive-going pulse must appear at U10-4 and a positive-going decode pulse must appear at U10-5 (top trace), and a positive (inverted) stop pulse must appear at U10-3. Scale: 5 milliseconds/centimeter horizontal, 2-volts/centimeter vertical.



Bottom trace is a pulse output from the character gate, U10-6, which triggers monostable U12. U12 output is shown in top trace. Normally U12-6 is low, as in center of top trace, but when triggered by the "good character" gate U10-6, the Q output of the character one-shot goes positive and remains positive, even into the next character (at machine speed). This action keeps the 100-μF memory capacitor charged. The expanded one-shot output is about 100 milliseconds long. This shows conditions during print. Pulse is only 40 milliseconds long during standby, so an occasional "good" pulse doesn't have much effect. Eight consecutive 40-millisecond pulses will turn on printer. Scale: 20 milliseconds/centimeter horizontal, 2-volts/centimeter vertical, unless otherwise noted.



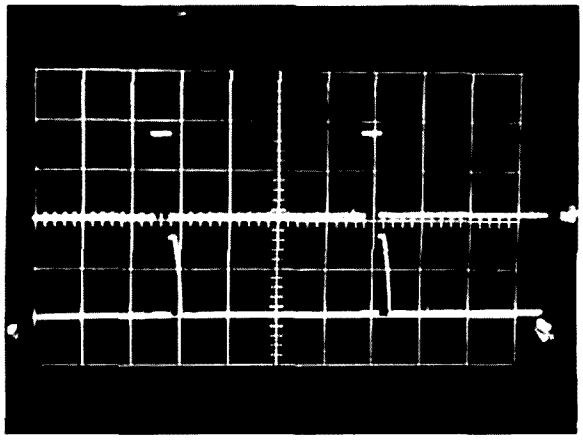
Top waveform shows output of character gate at U10-6 versus the incoming RTTY signal at U2-9. Location of this pulse in relation to the stop pulse is shown.



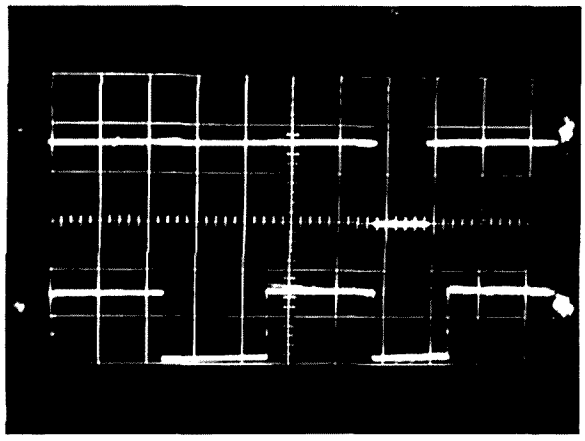
Top shows 4 clock pulses at pin 11 os 7474N divider. Bottom shows RTTY signal for letter I at input of digital TTL autostart (Q6 collector). Bottom os each trace is logic 0, or near zero Vdc.

nate the possibility of inadvertently shorting the high loop-supply voltage to the logic.

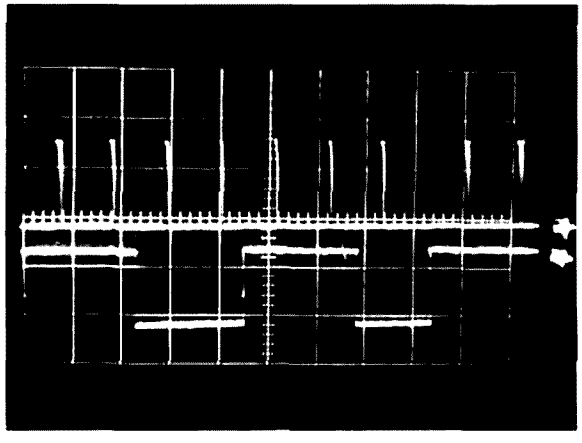
Construction is simplified and appearance enhanced if the available circuit



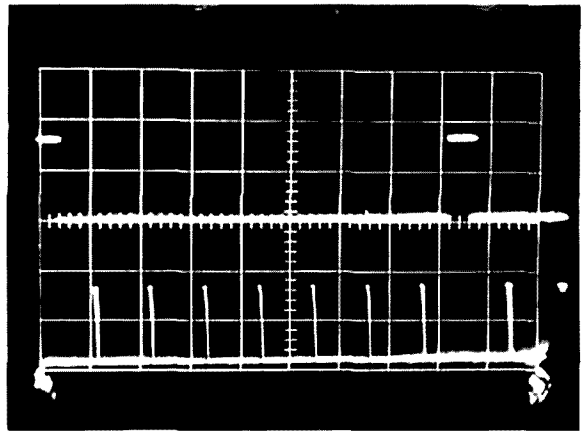
This shows shift pulse versus decode pulse (U9-8 and U13-6). Note relative timing. Positive-going edge of shift pulse (top) transfers RTTY character into storage, where it is checked 1.5 millisecond later by decode pulse action, which turns on the gates. Scale: 5 milliseconds/centimeter horizontal, 2-volts/-centimeter vertical.



Bottom trace is letter I sent at machine speed. Top trace is voltage at pin 9 of U8, the 7442N BCD decoder. This shows that the decoder goes low at pin 9 only during the stop pulse. This pulse is then inverted in U9 and enables both the anti-CW detector gate and the valid RTTY character gate, U10.



Top shows decode pulses (7 for each RTTY character). Bottom shows RTTY signal for letter I at input of autostart board (Q6 collector). Decode pulse checks each part of RTTY character at approximate center of the bit. Example: first positive-going part of square wave (bottom trace at left) is composed of start pulse (22 milliseconds) plus first bit (22 milliseconds), which is 44 milliseconds. Two decode pulses directly above this part of square wave show polarity of pulse and relative timing. The 7 decode pulses check start, plus 5 bits, plus stop pulse. Scale: 20 milliseconds/centimeter horizontal, 2-volts/centimeter vertical.



Bottom trace shows 7 decode pulses. Top trace is voltage on pin 12 of U2 start latch at machine speed. This voltage is used to externally sync scope during tests (negative slope triggering). Start latch goes off (high) during last portion of stop pulse until new start pulse triggers clock and generates new series of decode pulses at center of each 22-millisecond interval.

boards are used (see fig. 7). Each board is sent with a photograph that shows parts and jumper locations. Most of the component values are screened on the board.

critical components

The following suggestions are given to ensure proper operation. The small transistors should have moderately high h_{FE}. I selected mine from a large number of 2N706s with the aid of a scope curve tracer, but if you don't have a means of testing beta, use HEP-54 or MPS-2923, 24, or MPS3393 for Q1, Q2, Q3, Q6, Q7, Q8, Q9, Q10. Q11, and Q12. For Q4 and Q5, use pnp HEP-715, MPS-3702 or MPS-3703. These transistors are inexpensive and universally available. MJE340 transistors are recommended for Q13 and Q14.

The motor relay should have a coil resistance of not less than 120 ohms. Olsen no. SW-557 or Potter-Brumfield no. KA11DG dpdt, 12 Vdc, are satisfactory. (An almost identical relay constructed from the P & B relay kit is unsatisfactory due to too-low coil resistance.)

Pilot lights should be low current types, such as Chicago miniature

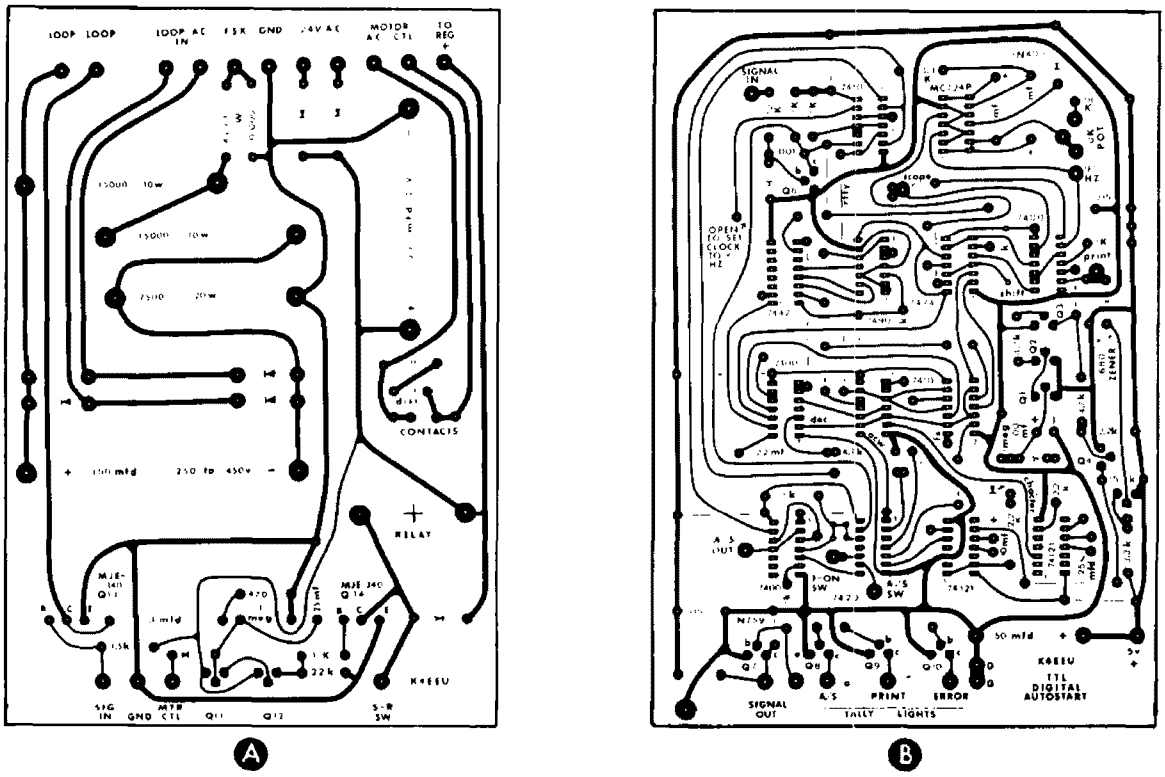


fig. 7. Etched circuit board layout for power-supply/motor control/loop, A, and autostart logic, B.

CM22-2-00-20, 6.3 V at 20 mA. HAL Devices also have suitable pilot lights if a different dc voltage is used.

The memory capacitor circuit is high impedance due to the long time constant. The 100-μF capacitor and the 1N914 diodes should have low leakage. Transistors Q1 and Q2 affect the loading, with the input resistance to Q1 approximately equal to the product of the betas of Q1 and Q2 and the resistance of the 4.7k base resistor to Q3.

The number of characters needed for turn-on is determined by the pulse lengths from the character one-shot, U12. The length of the pulse is set by an internal timing resistor at pin 9 of U12 plus the capacitor across U12-10, 11. In nonprint, the 3.3k resistor across Q5 emitter and collector is shorted out. Check the capacitor before installation to verify that it is close to 25 μF. More than 25 μF here will reduce the turn-on requirement and may cause false starts.

attachment to the demodulator

The ST-6 has a suggested sel-cal take-off point at pin 6 of the slicer op amp. If not already present, this additional connection may be made without affecting the operation of the ST-6. A phone jack may be added for a convenient external connection. The ST-5 is adapted similarly. The connection point should be the op-amp side of the 2.2k resistor The ST-3/4 connection point is the collector of pnp transistor Q2, with Q10 disabled so that the autostart circuit does not mute Q2. For other demodulators, the take-off point should be the processed signal in its best possible condition — not, for example, across the mark discriminator coil. The signal polarity should be as follows: The mark signal should be a positive voltage between 5 and 10 V; the space signal should be either zero or negative.

adjustment

Only one adjustment is required. The clock oscillator should be set at 91 Hz

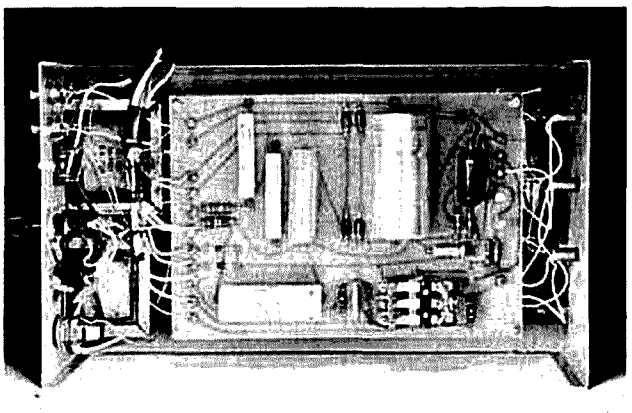
with a digital counter connected to a test point indicated on the printed-circuit board. Remove the jumper between U8-10 and U2-13, set the frequency, and replace the jumper. If all is well, the autostart should be in operating condition. When the receiver is tuned to the space frequency, both space and error tallies should light, and go off promptly when the receiver is retuned to mark.

The idle line turn-off time should be checked to ensure it is approximately 1 minute ±15 seconds. The voltage across the memory capacitor should be measured with a vtvm when checking becomes necessary. It will read about 2.8 volts during steady typing, about 1 volt on noise, and 1.9 volts will be the dividing line between print and nonprint. The turn-on character requirement should be seven or eight characters.

trouble shooting

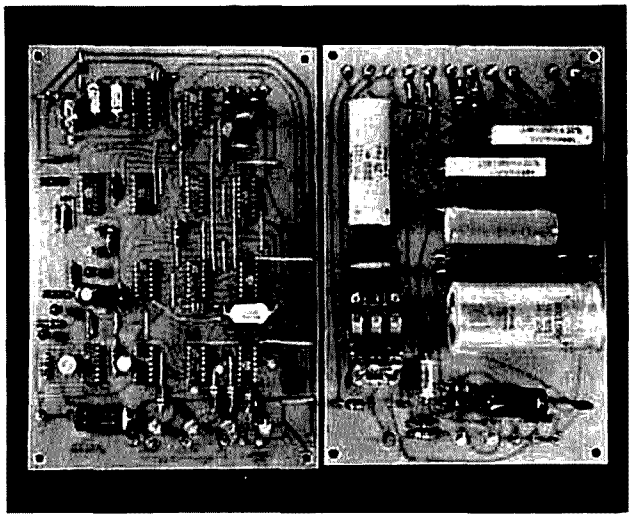
Trouble shooting is reasonably straightforward. A kit scope will usually suffice for signal checks through the unit. The scope should be synchronized on the test point provided to stabilize the trace and to make pulses appear in their relative positions on the trace. Typical waveforms of selected points in the circuit are shown.

With a ready-made circuit board, wiring errors are eliminated. Trouble spots to look for would be solder splashes, unsoldered connections, an IC installed backward, or a missing jumper.



Top view showing power-supply/motor-control/loop board installation.

If the autostart will not go into print unless the force-on switch is used, trouble in the logic is indicated. Isolate by disconnecting the three jumpers to the inputs of error adder U5, leaving only the trail connection to U5-1 and check to see if the error one-shot continues to fire. If so, the cause must be traced backward through U10-8. If the autostart begins to operate, connect the gates, one at a time, until the offending source of error pulses is isolated. Be sure that the test signal source is clean and that the clock is correctly set at 91 Hz. If all else fails, write me, giving results of checks and measurements made. Be specific, clearly state the difficulty, and enclose a self-addressed, stamped envelope.

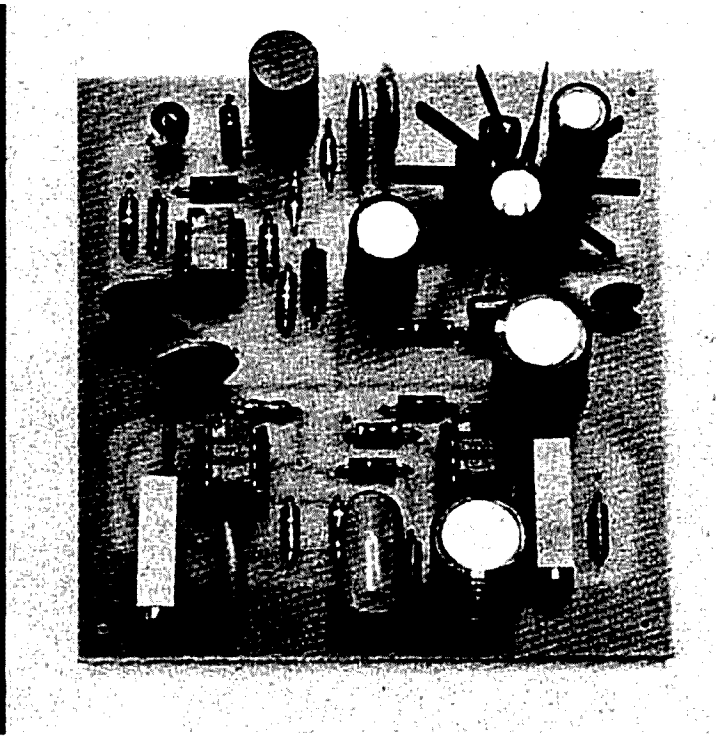


Logic and power supply/motor-control/loop boards.

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ham radio



a complete audio module

This audio module includes an active filter, preamplifier, audio compressor and power output stage

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When I recently decided to build a new receiver based on the modular concept, I spent several weeks delving into piles of magazines and clippings from trade journals to come up with the audio system described here. I don't know for certain why I started with the audio function block first, though on occasion my wife has accused me of doing things backwards. However, I am more inclined to believe I was influenced by the recent rapid advances in active-filter design and the abundant supply of circuits offered in the various publications. For whatever reason, the audio module came first, and since I felt some of the features might be of interest to others, I decided to write it up before going on to the next module.

features

Four distinct functions are provided in the audio package. These include a filter, a compressor, a preamp and a power amplifier. Both bandpass and band-reject functions are provided by the filter with bandwidth and frequency adjustable by

means of front panel controls. When not in use, the filter is bypassed.

The compressor, or audio agc amplifier, was considered to be a worthwhile addition to the complete system. It provides flat output with low distortion from about 20 mV to 2.0 V rms input. An external switch allows the compressor to be cut in or out, as desired. When the compressor is switched out, a preamp is automatically switched into the circuit to make up for the gain normally supplied by the compressor at low signal levels.

The final section uses a Motorola MC1454G power amplifier IC which is capable of providing up to 1-watt output for driving a speaker. A headphone jack is available for high-impedance headphones along with a switch to shut off the power amplifier if the speaker is not needed.

the filter

During my search through the magazines and clippings, many active-filter circuits were pulled and tried before I decided to use the circuit shown here.[1] Both bandwidth and frequency are independently adjustable by means of front-panel controls. With the values shown in the schematic of **fig. 2** frequency is variable from about 300 Hz to 2.0 kHz.

The original circuit was strictly a bandpass filter; a simplified circuit of which is shown in **fig. 1A**. In **fig. 1B** the large, hard-to-tune inductor has been synthesized by an RC network and op-amp IC, U2, resulting in a far more practical circuit. Since an extra op-amp was available on the board, it was used to implement the band-reject filter. In this mode, equal and in-phase signals are applied to both the inverting and non-inverting inputs of the op-amp, resulting in cancellation or zero output at the notch frequency.

Although the primary purpose for including a filter in the audio module was to aid in CW reception, the notch function can be quite useful for nulling heterodynes during phone reception. With the bandwidth control set at the sharpest position the 3-dB bandwidth is about 11 Hz.

Incidentally, a peaking function could

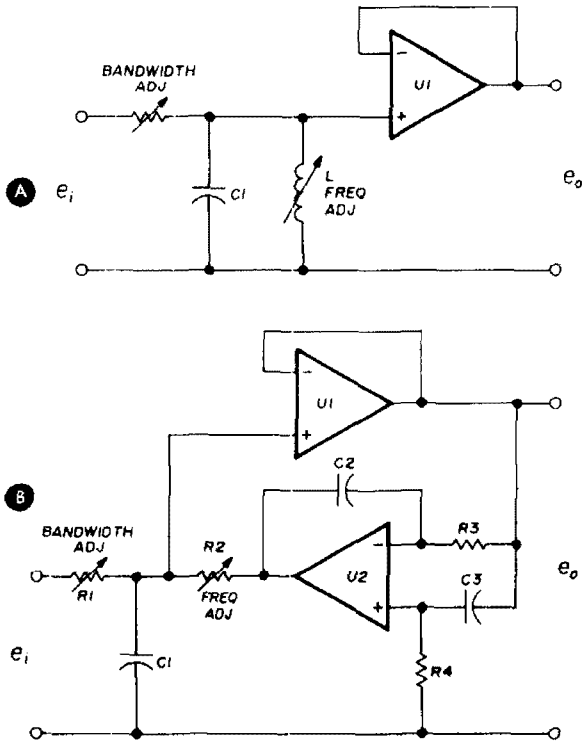


fig. 1. The basic active filter circuit used in the audio system. The simple bandpass configuration is shown in (A). Practical filter section is shown in (B).

be added to the filter if desired. Instead of using U2A as a subtracting amplifier, it could be used in the summing configuration with both signals fed to the inverting input. This would result in a 6-dB peak at the selected frequency. An additional position and one more pole would be required on the function switch to accommodate this extra function.

the compressor

Several audio agc circuits were tried, including one using the LM370 which provided squelch in addition to agc, but somehow I preferred this one over the others.[2] Referring to the schematic in **fig.**

2, you can see how simple the circuit is and how few components are required.

Use of the T-network in the feedback loop allows the circuit to handle signals considerably larger than in those cases where the fet appears directly across the tion to be adjusted when the compressor is switched out of the circuit.

The power amplifier uses an MC1454G IC and requires no additional comment since the circuit is straight from the data sheet.

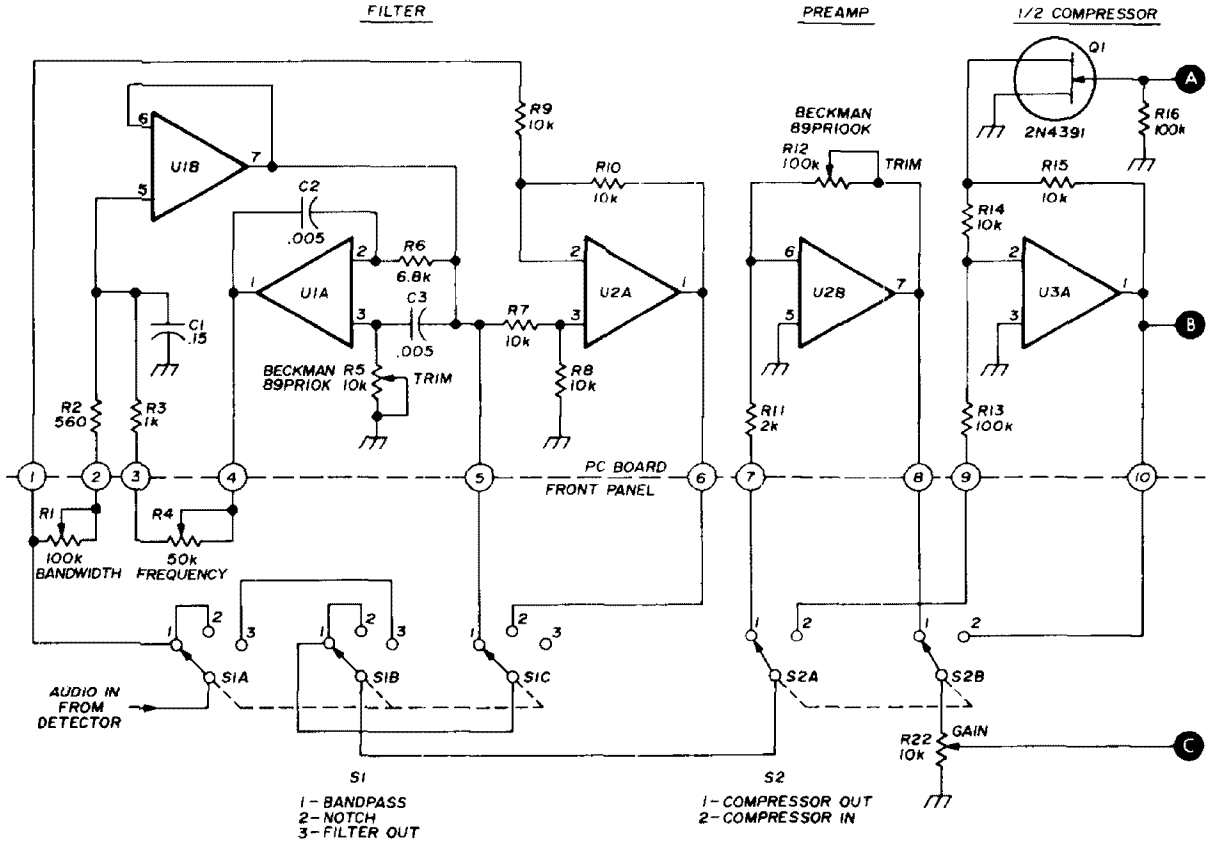


fig. 2. Circuit of the complete audio system module includes an active filter, preamplifier, compressor and power output stage.

signal path. Eventual gain of the circuit is limited by the maximum *on* resistance of the fet. The 2N4391 used here has a maximum of 30 ohms.

The voltage gain in my unit before compression begins is about 17. The agc has a fast attack characteristic with a slow decay. Electrolytic capacitor C5 controls the decay time and its value may be varied to suit your own requirements.

preamp and power amplifier

The remainder of the circuit is quite conventional, consisting of a standard preamplifier implemented with an op-amp whose gain may be adjusted by means of a trimmer resistor in the feedback loop. This permits gain compensa-

construction

Signetics type N5558V ICs were used for all op-amps on the board. These 8-pin dual-in-line packages contain two 741 op-amps each, allowing a very compact assembly. The board measures 3.2 by 3.4 inches and a layout is shown for those interested in duplicating the module. All external controls are connected by wire leads to holes provided in the board. If desired, the layout could be modified to suit a plug-in arrangement.

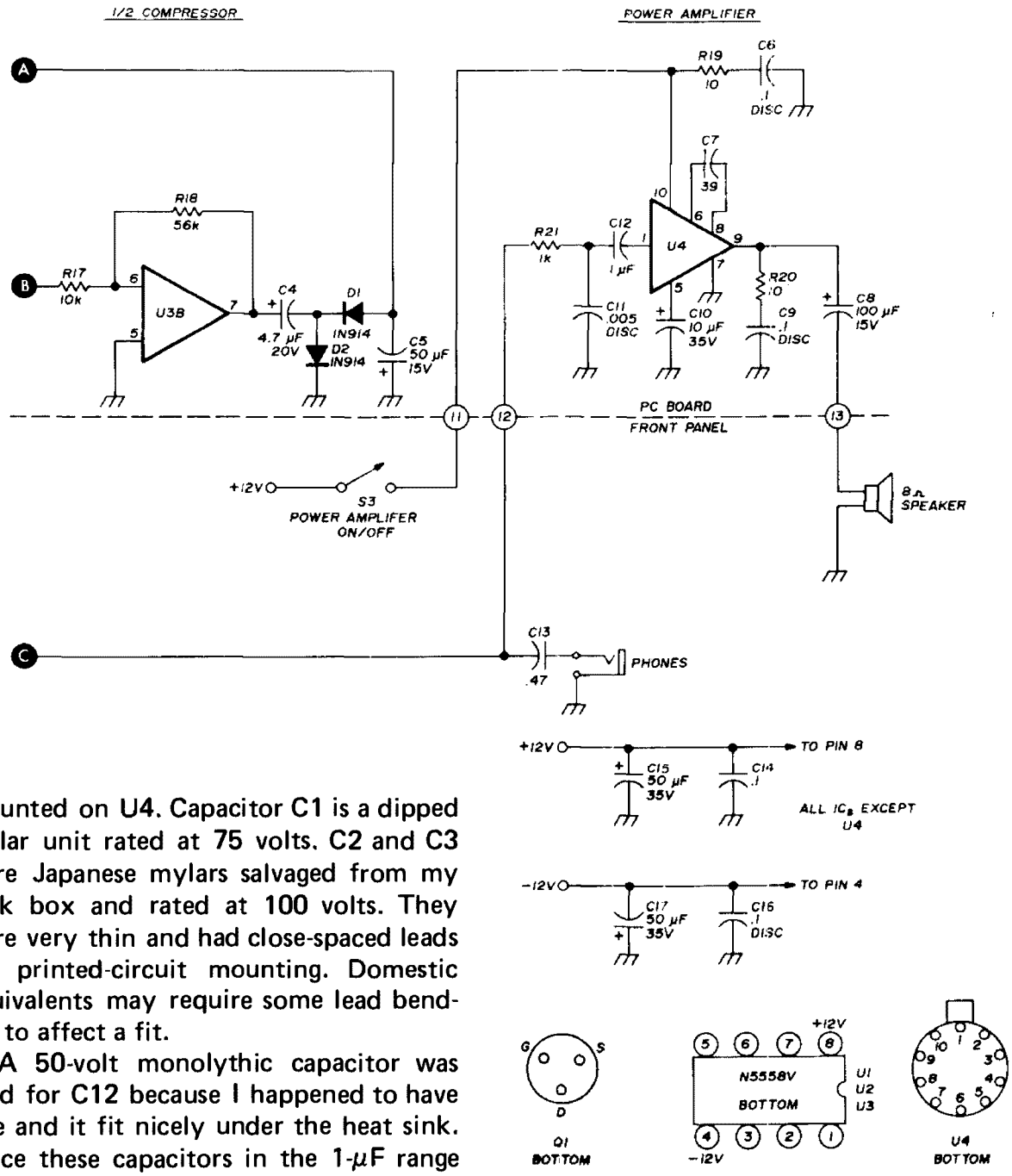
The following information may be helpful in obtaining parts physically suited to the hole spacings used in the original layout. All resistors are ¼-watt, 5%, though 10% units would be adequate. All the electrolytics are printed-circuit

types (single-ended) except C4, which is a small axial-lead tantalum.

The disc capacitors are 50-volt types and C7 is a dipped mica in the smallest size. This latter capacitor lies flat against the board so it will clear the heatsink

adjustments

A scope and an audio generator make filter adjustment a fast and simple procedure. Set trimmers R5 and R12 to mid-range before beginning. Switch out



mounted on U4. Capacitor C1 is a dipped mylar unit rated at 75 volts. C2 and C3 were Japanese mylars salvaged from my junk box and rated at 100 volts. They were very thin and had close-spaced leads for printed-circuit mounting. Domestic equivalents may require some lead bending to affect a fit.

A 50-volt monolythic capacitor was used for C12 because I happened to have one and it fit nicely under the heat sink. Since these capacitors in the 1-µF range are rather expensive, you may wish to substitute at the expense of a few mechanical problems. The 15-turn trimmer resistors are the small 3/4-inch units such as the Beckman 89P type used here. The 2N4391 in a metal can is around $3.00, depending on the manufacturer. The plastic version I used costs less than a dollar.

both the filter and the compressor and feed in enough signal at 1.5 kHz to obtain an output of 1 or 2 volts p-p as seen on the scope when connected to the high end of the gain control.

Set the bandwidth to maximum resist-

ance (sharp) and turn the function switch to *bandpass*. Carefully adjust the frequency control for a peak. Trimmer R5 should then be adjusted for an output amplitude equal to that obtained without the filter. Make sure you're right on frequency, it's quite sharp! That's all there is to it.

The other trimmer, R12, controls the gain of the preamp and may be adjusted to suit your own requirements. Clockwise rotation of the slotted shaft increases gain. The compressor requires no adjustment, but while the scope is connected you can check the circuit for proper operation. With the compressor switched in, output should be around 1.0 V p-p with input levels from 60 mV p-p to around 6.0 V p-p. Of course, there may be some differences due to variations in the characteristics and tolerances of the components.

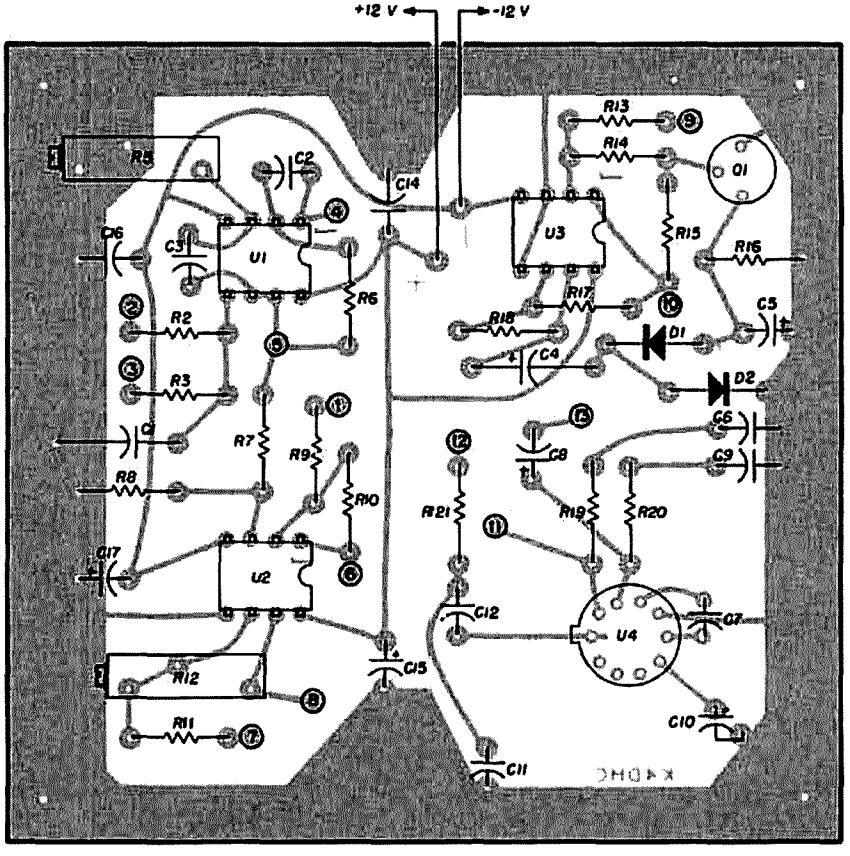


fig. 3. Full-size layout of the component side of the printed-circuit board. Numbers correspond to those used in fig. 2 for external connections.

operating hints

The active filter is quite sharp and takes some getting used to. When using the notch function, turn the bandwidth control to minimum resistance (broad) or you may never find the null. In the bandpass mode, the broad bandwidth setting can be used for phone reception since rolloff is quite gradual.

As I have done in the past with several other articles, I shall attempt to help out with any of the more unusual parts if it is necessary. In this case the Beckman trimmers and the plastic 2N4391 are about the only items which may be troublesome. Drop me a line and a self-addressed, stamped envelope if you need help.

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ham radio

fm repeater installation

Suggestions on dealing with setup and control problems

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A recent article in *ham radio*[1] caused many questions on how to set up a repeater and how to solve the problems encountered. These notes are presented to help answer these questions and help solve some of the problems.

definition of a repeater

A repeater is an fm, TV, facsimile, or similar station that receives a signal on some input frequency and automatically transmits the received signal on some output frequency. The purpose of a repeater is to extend the communication range between a group of stations. The repeater generally consists of a receiver with its antenna, a control unit, and a transmitter with its antenna. This simplest form of an amateur fm repeater station is shown in **fig. 1.**

setting up a repeater

From the block diagram of **fig. 1,** it is seen that a basic repeater station consists of five components: two antennas, a receiver, transmitter, and control unit. When setting up a repeater, care must be taken to select each component carefully so that the components will function as a reliable repeater system rather than a collection of pieces of equipment with a high maintenance rate and low performance. For example, when selecting a receiver, care must be given to its selectivity, sensitivity, and intermodulation characteristics.

The receiver must be sensitive enough to receive low-power signals within the desired coverage area; on the other hand it must not be so sensitive as to pick up signals greatly removed from the desired coverage area. The receiver must be selective enough to eliminate strong signals on adjacent channels, but not so selective that most signals are too wide for the receiver passband. The receiver must be relatively insensitive to intermodulation products. For example, if two commercial stations are on the air simultaneously from a nearby site, the receiver must not mix the two signals, causing the reception of unwanted signals.

The transmitter must have enough power when used with the transmitting antenna to produce an adequate signal

within the desired coverage, but must not be so powerful that it causes interference in outlying areas. The transmitter, furthermore, must generate a signal free from spurious radiation and transmitter noise. The antennas should have sufficient gain for adequate receiving and transmitting coverage within the desired area. Each item in the repeater should be chosen to be as maintenance free as possible. This does not exclude tube equipment, but it does mean that the equipment must be capable of operating continuously at some remote location without frequent maintenance.

hooking it all together

Once the receiver, transmitter, and antennas have been assembled, the only remaining task is to connect the equipment together and see if it will work. To do this, it will be necessary to build some type of carrier-operated relay circuit, which will turn the transmitter on when a received signal is present. Fig. 2 shows a simple Carrier-Operated Relay (COR) circuit for use with a tube receiver. This circuit is similar to the COR used in reference 2, which should be consulted for connection to various types of receivers. For a very simple repeater, the relay itself can be used to key the transmitter directly when a signal on the input frequency is received. While this method will work, direct transmitter keying has some very serious disadvantages. For one, if something goes wrong and the transmitter suddenly decides to generate a birdie on the input frequency, the repeater will latch up and remain in operation even though an input signal is not present. In addition, some long-winded fellows could keep the repeater keyed for days at a time without giving the other stations a chance to talk.

Another problem with this simple circuit is that the transmitter will undergo rapid on-off cycles during periods of short transmissions. This will produce excessive wear and tear on mechanical relays, and the resulting series of double squelch crashes is hard on the ears. These problems can be cured by inserting a reliable control unit between the receiver and the transmitter. This control unit will "time out" any received signal that is on longer than some predetermined interval (say, three minutes) and will provide a carrier dropout delay so that the transmitter carrier will not drop out between transmissions. The hookup of this control unit is shown in fig. 3. The control unit is described in reference 1. Note that fig. 3 also shows a relay in series with the transmitter keying lead. This relay is used for turning the repeater on and off either locally or via remote means such as a leased telephone line (radio line).

Note also that some additional items will be needed for a legal repeater. An

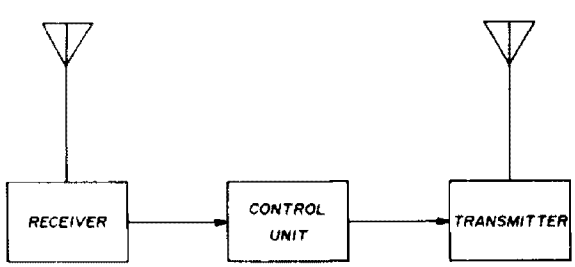


fig. 1. Basic repeater setup.

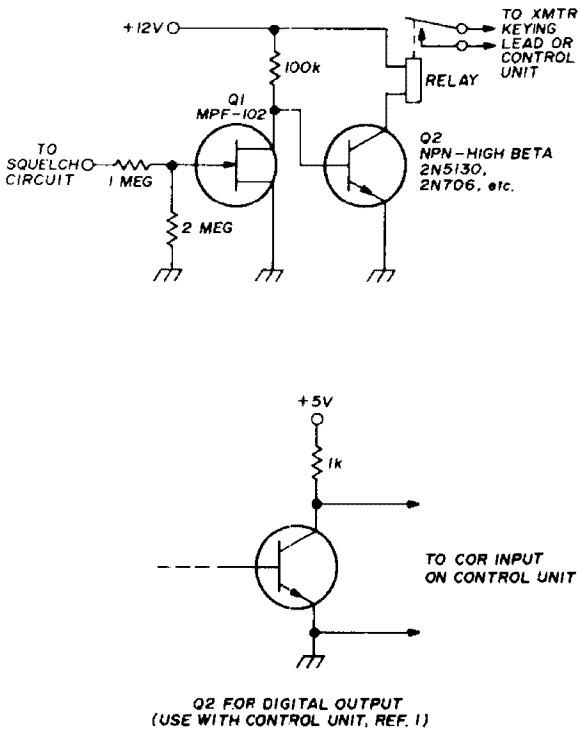


fig. 2. Simple carrier-operated relay circuit.

identifier of some sort will be required to identify the repeater automatically at five-minute intervals, either by phone or CW. The referenced repeater control unit provides a timer that will trigger the identifier as long as the repeater is active. Another item required by law is a monitor to monitor the input and output frequencies.

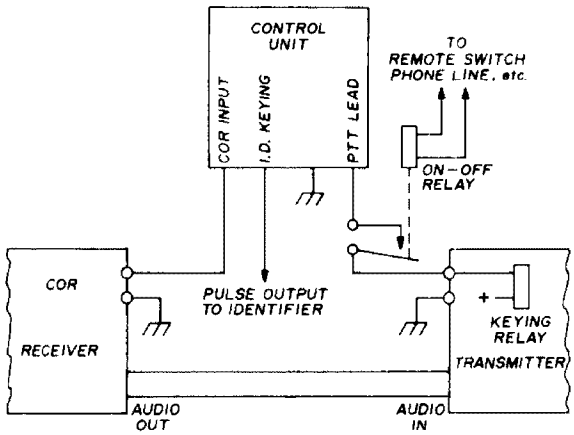


fig. 3. Typical control-unit hookup.

some problems involved

While it is possible to hook a repeater together as previously described and have it work the first time, few people are so lucky. Most repeater groups encounter problems of one type or another.

One of the most common problems is desensitization. This problem is caused by too much transmitter power getting into the receiver front end. This power will desensitize or block the receiver. The result is that the receiver will only respond to very strong signals, or to no signals at all, when the transmitter is operating. This problem can be identified very easily by listening to a weak received signal and monitoring the first limiter with a meter (transmitter off). If, when turning on the transmitter, the signal gets weaker or disappears, and the first limiter current goes *down*, you have a desensitization problem. If the signal gets weaker, and the first limiter current goes *up*, the problem is caused by transmitter-generated noise.

Desensitization can be cured in several ways. The transmitter and receiver should be well shielded, and power leads from inside the shielded units should be well bypassed. The antennas should have a large vertical separation to prevent transmitter power from getting back into the receiver. Most commercial two-way radio dealers have charts giving the separation required for frequency versus amount of isolation desired.

The problem of transmitter noise is another matter. If a commercial, type-accepted unit is used as the transmitter, it is unlikely that transmitter noise will be a problem. If home-brew equipment is used, transmitter noise is likely to be an annoying problem. Noise can be eliminated either by redesigning the transmitter output stages or by installing one or more cavity filters in the transmitter output. The cavity filter is very selective and can be successfully used to eliminate transmitter noise from the receiver input. The best way to use the cavity is shown in fig. 4. In this case, the cavity is inserted in the transmitter output lead and is used in a suck-out mode. In other words, it passes all rf energy but sucks-out the energy or noise at the receiver input frequency. Some of the references given explain the proper use of cavity filters.

Cavities can also be used to help reduce desensitization, but don't get the idea that cavities are a requirement in a repeater. They are not. As an example, the Liverpool repeater, in Liverpool, New York, operated for over a year without cavities and without desensitization or transmitter noise problems.

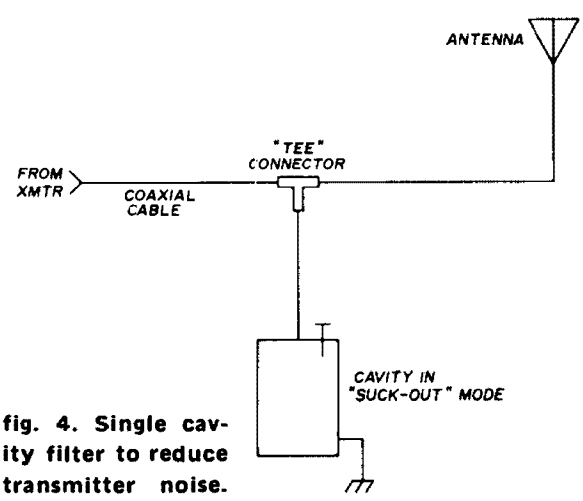


fig. 4. Single cavity filter to reduce transmitter noise.

choice of frequencies

While there is no rule giving a list of input and output frequencies and channel separation for repeaters, it is recommended that the current plan suggested by the ARRL be used. If this plan is supported nationally, it will be possible to travel coast-to-coast and work into a large number of repeaters using only a handful of crystals.

This article isn't intended as a short course on repeaters, but as a guide for dealing with some of the problems that may be encountered in repeater installation. The articles in the references provide essential details and should be read before attempting to set up your repeater.

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ham radio

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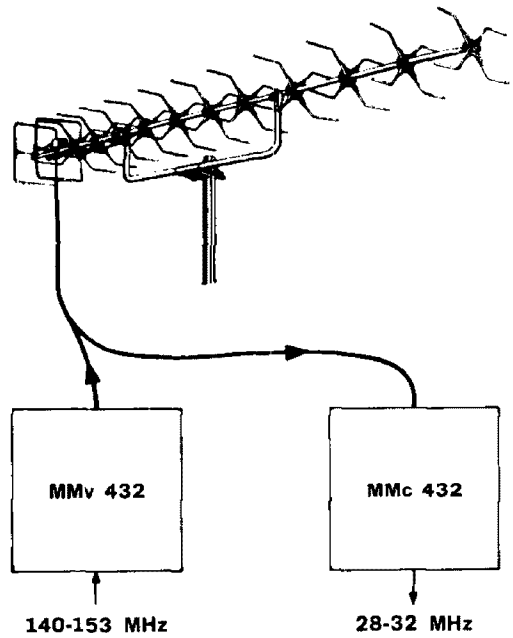
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regulated ac power supply for mobile fm equipment

This low-cost regulated power supply provides a convenient way of powering your mobile vhf fm equipment at your home station

Paul J. Dobosz, WA8TMP, 855 East Commerce Road, Milford, Michigan 48042

If you have been looking for a good regulated 12 to 14-volt dc power supply to use your mobile vhf-fm solid-state gear at home, look no further. This simple low-cost regulated supply will power most solid-state rigs and a beefed up version will power most of the popular rf power amplifiers currently available for mobile use. Power supplies suitable for this purpose are available from electronic suppliers and transceiver manufacturers, but the prices are far from cheap. That fact makes construction of a supply the only economical alternative.

The design is about as simple, versatile and straightforward as a regulated power supply can get. It can be tailored to suit a wide variety of individual needs from milliwatt HT to high-power linear. The best news is the price. The basic 3-amp version costs slightly less than $25, using brand new parts. A 6-amp model can be built for an additional $5. Regulation is between 5 to 8 percent for full load with ripple of not more than 100 millivolts.

the circuit

Transformer selection is not critical as long as the transformer meets these simple requirements. It must provide the necessary secondary current at 15 to 20 volts rms.

The full-wave bridge rectifier can consist of four individual diodes with a PIV of 50 volts or greater, or an epoxy encapsulated bridge can be used. Current capacity of the individual diodes or bridge should be equal to (or preferably, greater than) the amount of current to be drawn from the supply. The use of a pre-packaged bridge is advantageous because of its low cost, compactness and convenience. Individual diodes might prove more economical though if they are readily available.

The capacitive filter consists of a single electrolytic capacitor connected across the output of the bridge rectifier. The only requirement for the capacitor is a capacitance of at least 2000 μF at 35 to 50 volts dc. A smaller amount of capacitance will work but the ripple will increase considerably as the value of capacitance is decreased. If the ripple reaches more than 500 millivolts it will begin to appear on the transmitted signal as hum.

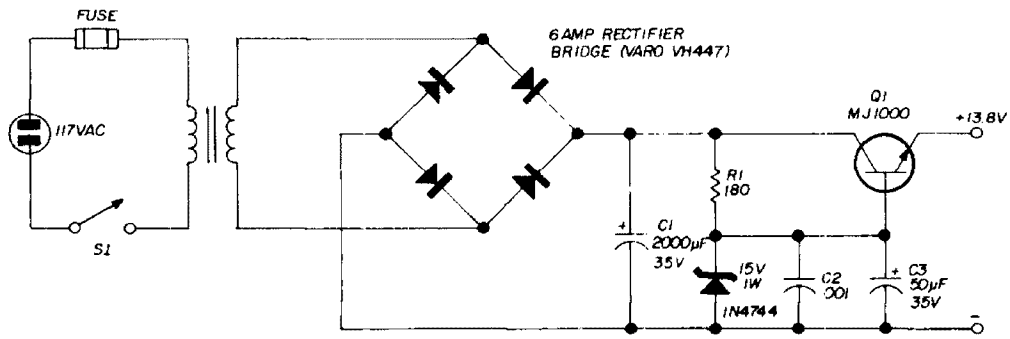


fig. 1. Circuit diagram for the regulated low-voltage power supply. Transformer T1 is a Triad F47U (3 amps) or Triad F48U (6 amps). For current capacity greater than 3 amps, transistor Q1 should be paralleled with another MJ1000 for each additional 3 amps of output current.

regulator

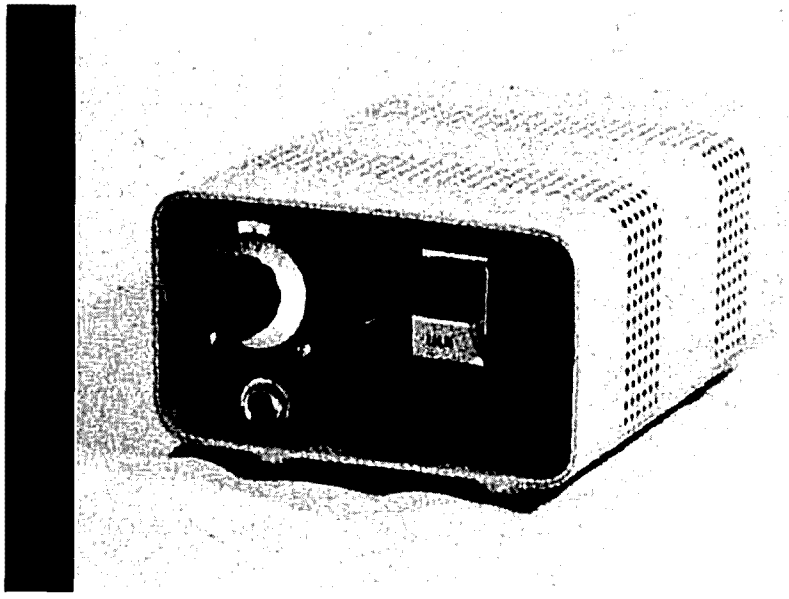
The regulator circuit is a simple series regulator using a zener diode as a voltage reference. The heart of the regulator is the series regulator transistor which is a pre-packaged Darlington amplifier with extremely high gain. This allows the use of inexpensive 1-watt instrument zeners as the voltage reference. The high beta of the MJ1000 transistor means that only one or two milliamps of base current are required to provide each amp of output current. The MJ1000 is rated at 5 amps but it is necessary to reduce the load of the MJ1000 to about 3 to 4 amps per unit because of the power dissipation rating of the device.

This conservative design will prevent thermal damage to the MJ1000. If more than 3-amp capacity is needed, the additional current can be obtained by paralelling MJ1000s along with increases in the capacity of the bridge rectifier and power transformer. Selection of the zener diode voltage should be approximately 1.2 volts higher than the desired output voltage to overcome the two base-to-emitter drops of 0.6 volt each in the Darlington.

construction

The power supply can be constructed on almost any chassis or cabinet. Although it is not absolutely necessary it would be a good idea to fuse both the ac input and dc output of the supply. The MJ1000 should be heat sinked and mounted in the clear so that air can circulate freely. If an epoxy encapsulated bridge rectifier is used it should also be heatsinked. All wiring should be at least no. 18. Layout is not critical so your imagination is the only limit for a custom installation.

ham radio



micropower communications receiver

This high-performance solid-state 20-meter receiver requires less than 80 milliwatts of power

Michael J. Gordon, Jr., WB9FHC, 1615 Elmwood, Wilmette, Illinois 60091

If you've ever tried to operate a typical transistorized receiver from dry-cell batteries you know how expensive it can be. The current drain of most receivers is on the order of hundreds of milliamperes and battery life is short. For example, a General Electric fm transceiver draws 500 mA in the receive mode, with the transmitter section off and the final tube filaments disconnected. At twelve volts this represents total power dissipation of 6 watts, making dry-cell operation impractical. Of course, this is a rather complex receiver and, in that light, this may not be considered an excessive amount of current drain.

However, consider a much more simple receiver such as the *beginners receiver* described in the 1970 ARRL Handbook. This is a direct-conversion receiver and uses only two transistors and a single IC; yet, current drain is 24 mA and power dissipation is about 350 milliwatts.

In the receiver presented in this article the total power dissipation is only 80 milliwatts. Unlike the simple direct-conversion receiver it features performance superior to the receivers in a great many commercial transceivers at least in regard to sensitivity, noise figure and stability.

Selectivity is determined by how much you are willing to pay for a crystal filter, and in this respect a transceiver may be slightly superior, although no significant differences may be noted in use. Also, this receiver covers only one band, twenty meters, and is designed primarily for CW operation.

A low-power receiver is especially advantageous in an emergency situation when the only source of power is a battery. With the low current required by this unit, many hours of operation can be

obtained from a single set of batteries. For example, a few surplus sub-C size nicad batteries rated at 750 mA hours will provide more than 60 hours of continuous operation. And, even during non-emergency situations there are advantages to a low power, high performance receiver. When operating portable, for example, a few AA batteries will provide many hours of operation and weigh very little.

used here because of its lower cost compared with a fet. After the signal passes through the crystal filter, cross modulation is no longer a problem so other considerations were given more priority. The last i-f amplifier is coupled to the detector, an RCA CA3028A IC. This IC is well suited for this application and the circuit has extremely high conversion gain and good overall performance as a product detector.[1]

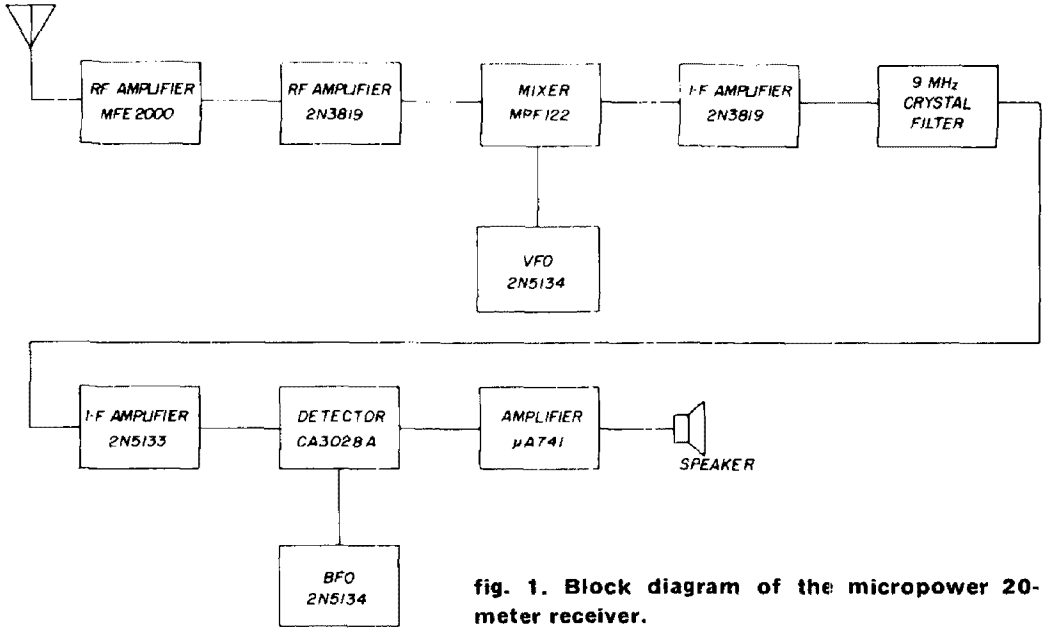


fig. 1. Block diagram of the micropower 20-meter receiver.

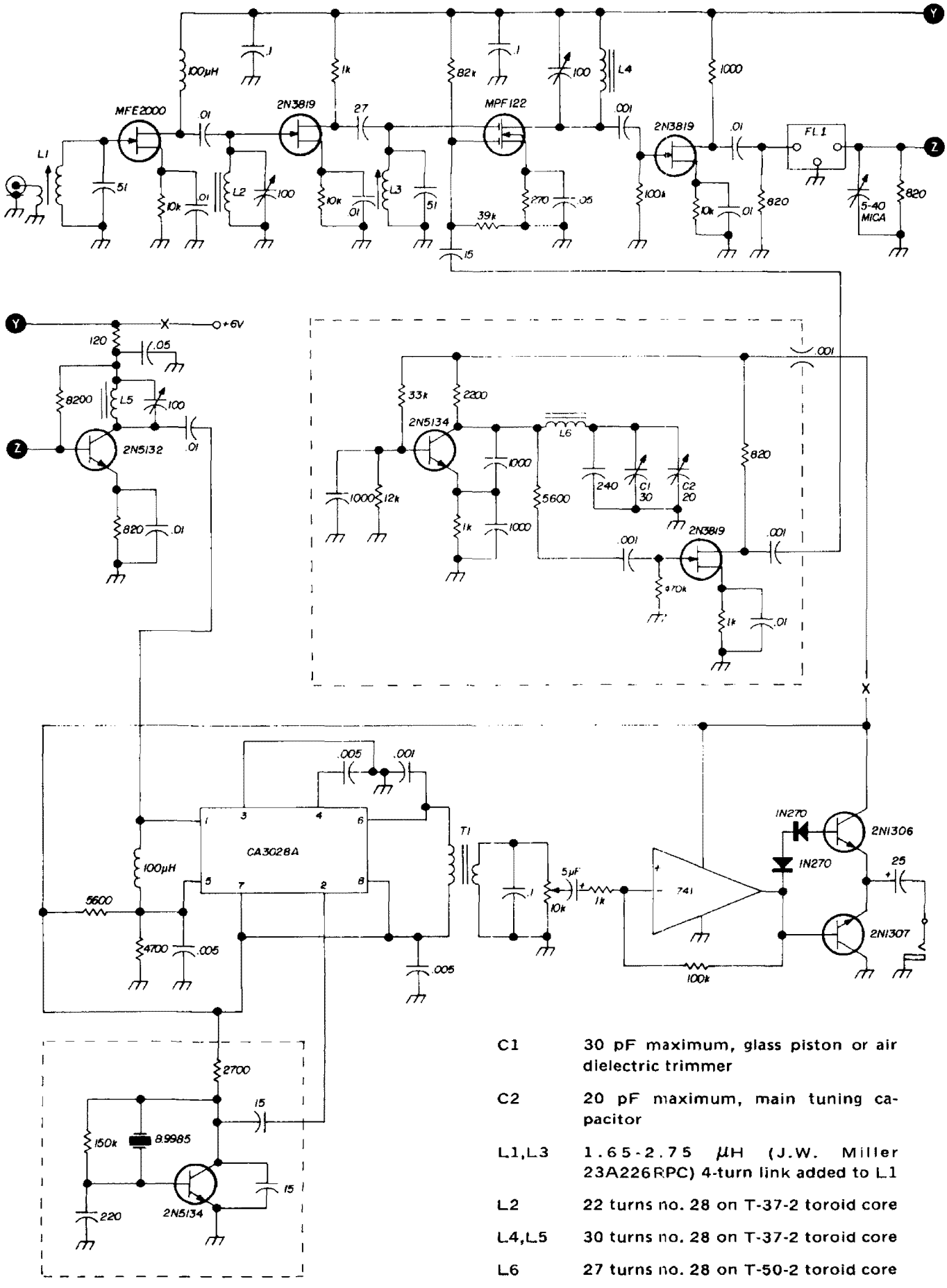
circuit

A block diagram of the complete low-power receiver is shown in **fig. 1.** A signal at the antenna is amplified by two fet rf stages and fed to a mosfet mixer. The MPF122 mosfet is a low-cost plastic device and features diode protected gates. The 9-MHz output of the mixer is amplified by another fet which also acts as an impedance matcher, transforming the high-impedance output of the mixer to the lower input impedance of the crystal filter.

After filtering, the signal is amplified by a high-gain bipolar transistor. The base-bias network of this transistor serves to properly terminate this side of the crystal filter. A bipolar transistor was

The output of the detector is fed to the audio gain control and from there to the audio amplifier, a simple circuit using an IC op amp and a pair of complementary transistors. The bfo is a single transistor in a simple crystal oscillator circuit; the vfo is a two-transistor circuit. One transistor is in a Vackaar oscillator circuit, the other, an fet, is used as a buffer.

The two rf amplifiers are almost identical, with the exception of the method in which the supply voltage is fed to the drain. In one case an rf choke is used for minimum voltage drop, and in the other a resistor is used. Two stages of rf amplification are used because the gain of a single stage operated at these voltage and current levels is not adequate for the



C1	30 pF maximum, glass piston or air dielectric trimmer
C2	20 pF maximum, main tuning capacitor
L1,L3	1.65-2.75 μH (J.W. Miller 23A226RPC) 4-turn link added to L1
L2	22 turns no. 28 on T-37-2 toroid core
L4,L5	30 turns no. 28 on T-37-2 toroid core
L6	27 turns no. 28 on T-50-2 toroid core
T1	10k:2k miniature audio transformer

fig. 2. Complete circuit diagram of the micropower receiver. Total current drain with a 6-volt power supply is about 15 mA. The crystal filter is a Hermes X9MA or equivalent.

performance I wanted. Substitution of the MFE2000 is not recommended unless you replace it with a device of similar specifications. Even the 2N3819 produced noticeably poorer performance as far as noise figure and tendency to oscillate are concerned.

The mixer is a standard circuit, and the dual-gate mosfet features excellent conversion gain as well as diode-protected gates and low cost. Gate two is biased at one-third of the supply voltage to increase conversion gain. Oscillator injection is through a 15-pF mica capacitor to minimize pulling of the vfo, whose output is a very stable signal, nominally in the 5.0- to 5.150-MHz frequency range.

The mixer takes the difference between the incoming 14-MHz signal and the 5-MHz vfo to produce the 9-MHz i-f, which is further amplified. The coil in the vfo is a toroid which provides a high-Q inductance in a minimum amount of space. Vfo stability is excellent, partly due to this high-Q coil, and partly because the transistor parasitic capacitances are swamped by external capacitors on all elements. These capacitors must be high quality silver-mica units if any degree of stability is desired.

The first i-f amp is fairly straightforward and it is coupled to the crystal filter through a capacitor. The crystal filter determines the selectivity of the receiver, and if substitutions are made, there are a few things to keep in mind. First, the circuit shown was designed around a surplus four-crystal unit which had no dc path between the output pin and ground. In order for this circuit to work there must be no dc path through the filter between the base of the second i-f amplifier transistor and ground. If there is, the base bias will be upset and the circuit will not function properly.

If your filter has a dc short from the output to ground, as measured on an ohm-meter, simply insert a .01-μF capacitor between the base of the transistor and the filter. The base-bias network current is set to allow for proper impedance termination of the crystal filter specified.

The output of the second i-f stage goes to the product detector which is a circuit from the ARRL Handbook. Although it was not intended to be operated from a 6-volt supply, it seems to work fine at that voltage, having high gain and a good output level.

The output of the product detector is transformer coupled to the audio amplifier, which consists of an IC op amp driving a pair of transistors operating in Class B. The current drain of this circuit is extremely low and the amplifier is very efficient. Since there is no provision for avc, and since the receiver is designed primarily for CW reception, the audio amplifier is designed to clip at a certain input level above which the audio output cannot rise. Although this produces severe distortion on strong signals, the ease of copying the CW is in no way affected. This provides a very effective means of limiting the dynamic range of the output volume without impairing performance. In fact, it represents a marked improvement over the receiver with no avc at all.

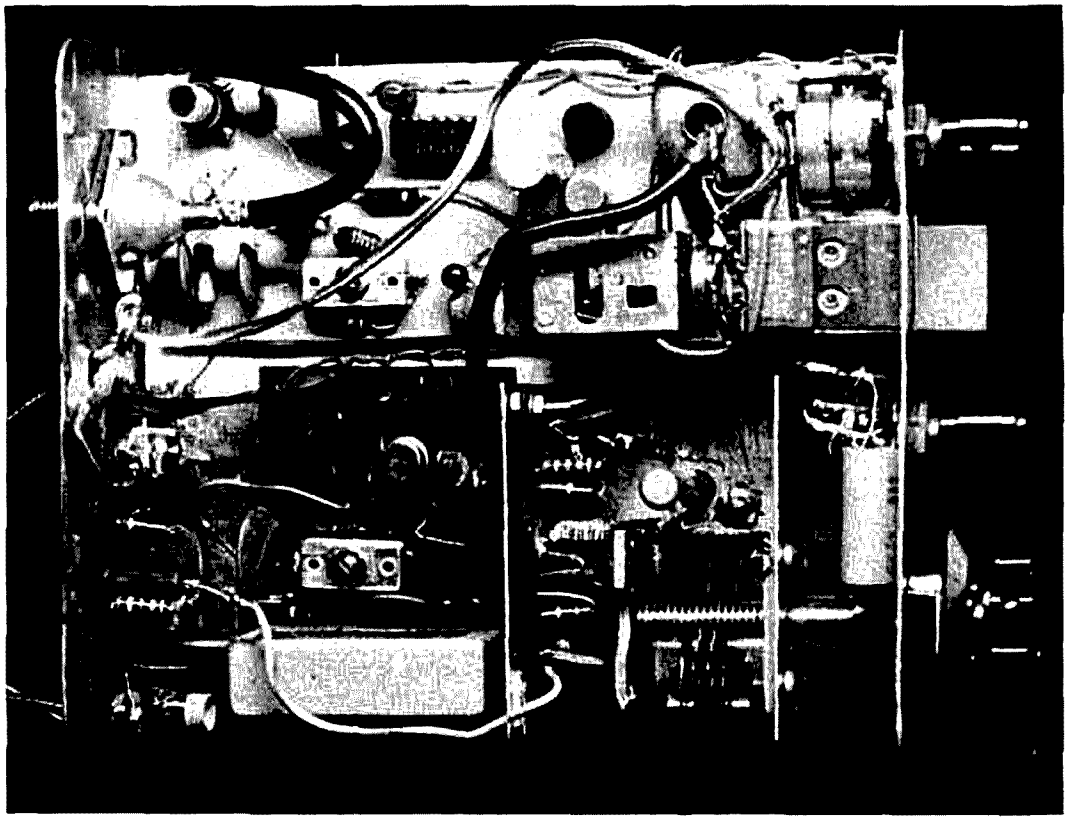
The output of the audio amplifier is sufficient to drive a pair of sensitive high-fidelity type headphones to adequate volume, if the elements are wired in series. However, the recommended load is a good pair of sensitive 600-ohm headphones. It is important that the headphones shut out as much external noise as possible since the audio output is not great. The output coupling capacitor was selected with this headphone impedance in mind. With lower impedance phones you can expect a loss of low-frequency response.

construction

A modular type of construction was used, with one or two stages in the receiver on each of several printed-circuit boards. All of these boards were made by hand, using nail polish for resist. This method proved to be adequate for circuitry of this type where there are no really tight foil runs. This method is very good because it allows for change in the circuitry as deficiencies show up during construction, without necessarily having

to go through a multi-step process of preparing a photographic pattern, exposing the board, and the rest of the work that goes with making a photographically etched circuit board. This is not to slight the photographic process, which is great once you have the breadboard version of your project working and are ready to make a *professional* version.*

The vfo and bfo were put into individual compartments for shielding, the former in a small minibox, the latter in an old i-f transformer can. Mechanical stability of the vfo is an important consideration and no effort should be spared to achieve it. The minibox mounted on another chassis proved to be adequate in this case. Everything was mounted in a LMB CO-3 cabinet which presents a fine looking piece of equipment when complete. A low-cost vernier dial was used for tuning purposes, and it works well even though you are not able to read frequency directly from its calibrations.



Construction of the micropower receiver.

alignment

Alignment is simple, requiring nothing other than a calibrated receiver or transmitter for reference. The vfo is set very simply by placing the main tuning capacitor at maximum capacitance and adjusting the trimmer until the received frequency is that frequency at which you wish the lower band edge to fall.

In view of the fact that the vernier dial will give you only an approximation of what frequency the receiver is tuned to, it is wise for advanced and general-class licensees to set the lower band edge a kilohertz or so above 14.025 MHz so you cannot hear, and therefore will not be tempted to transmit to, any stations outside of your frequency segment. By the same token, the Extra-class licensee should set the low end of the band at roughly 14.001 MHz.

It is not critical where the high end of

*A complete etched and drilled PC board is available from Psynexus Systems, 445 Oakdale, Glencoe, Illinois 60022 for $3.50. A complete parts package is $60.00.

the band falls so no further adjustments are necessary. You should find that the high end of the band is roughly 150 kHz above the low end. If you find it necessary to change the bandspread, the following information will be useful. First, decreasing the inductance of the vfo coil by removing turns and then adding compensating capacitance in parallel with the main tuning capacitor will increase bandspread. That is, the tuning range will be made smaller. Conversely, increasing inductance by compressing the wire on the coil or adding turns, and decreasing capacitance to compensate will decrease band-spread (a greater range of frequencies will be covered).

Rf and i-f alignment is simply a matter of peaking the various coils for maximum signal strength. A fairly weak signal should be used, and the audio amplifier should be kept below the limiting point by keeping the gain as low as possible. You may wish to temporarily hook up an external audio amplifier for these tests in order to provide more volume. When the unit is properly tuned you will notice a significant increase in the output noise level when your 20-meter dipole is connected up to the antenna terminals. When the band is open, a 3- or 4-foot piece of wire will bring in plenty of signals.

conclusion

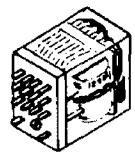
All in all, this receiver makes an excellent construction project. The circuit is simple enough for the average ham with a little previous experience to build, yet performance is superior to many commercial transceivers on the market. Cost is reasonable, and the receiver is better insofar as current drain is concerned than anything else available. All you need is a simple QRP transmitter, such as the Ten-Tec TX-1 to make a complete high efficiency amateur station, ready for any portable or emergency situation that may come up.

reference

1. *The Radio Amateurs Handbook*, A.R.R.L., 1970, pages 137-142.

ham radio

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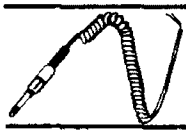


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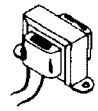


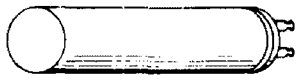
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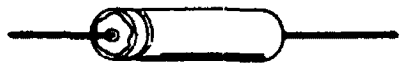


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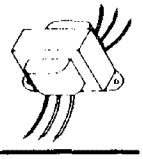
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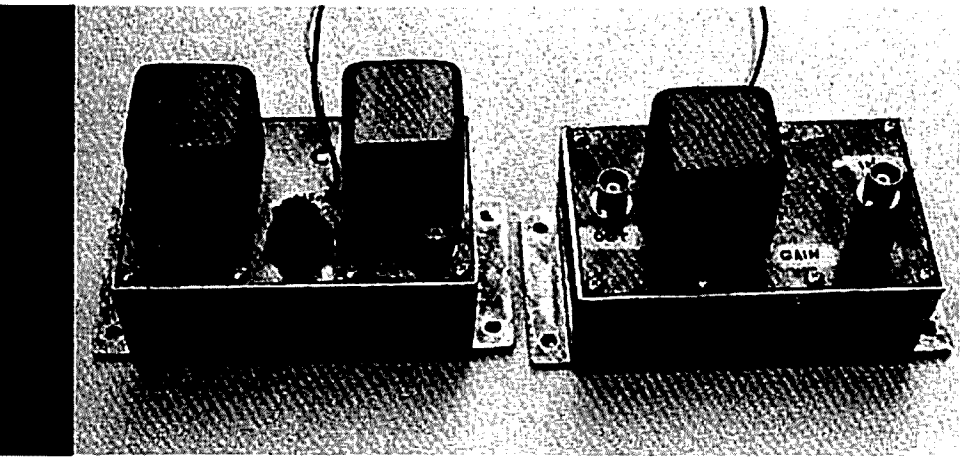
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high-performance broadband ic amplifiers

Construction details for two broadband IC amplifiers which provide up to 40-dB gain over the frequency range from 1 to 56 MHz

Hank Olson, W6GXN, Post Box 339, Menlo Park, California

Presented here are a pair of IC amplifiers which are useful in a variety of measurement and communication systems problems that lie between 1 MHz and 56 MHz. As such, they encompass the seven lowest amateur bands, providing useful gain at these frequencies.

Both amplifiers make use of the Motorola MC1590G integrated circuit which is characterized as an i-f/rf amplifier by the manufacturer. The more simple of the two amplifiers, fig. 1, uses only one broadband 400:50-ohm ferrite transformer, providing approximately 25-dB gain and a noise figure of about 13 dB at 30 MHz. The more complicated amplifier, fig. 2, uses two broadband transformers and provides approximately 40-dB gain and a noise figure of about 3 dB at 30 MHz. Both amplifiers have a gain control.

The quoted noise figures for both amplifiers were measured at maximum gain with an AIL Model 75 automatic noise-figure meter. Since the typical noise figure for the MC1590G integrated circuit is the only one specified (6dB at 60

MHz), it is possible that somewhat different values be obtained with individual ICs.

applications

Just what good is a broadband amplifier to the amateur world? Probably the same question was asked about the broadband double-balanced hot-carrier mixer when it first arrived on the scene. Since hams are usually concerned with bands that represent only a few percent bandwidth, broadband devices are not a requirement. However, the convenience of having a broadband device that can be used on *any* of the lowest frequency ham bands, without switching, tuning or adjustment is a real asset in rf measurements or when a system is being first checked for concept.

Consider, for example, that you have just heard on the news that the Russians have launched a satellite that transmits on 40 MHz – complete with dog heart beats. You quickly jury-rig a receiver to hear the satellite transmissions. A signal generator is used for a local oscillator, a double-balanced hot-carrier diode mixer is used to convert 40 MHz to a convenient i-f, say 7 MHz and an amateur-band receiver is used as an i-f amplifier, detector and audio amplifier. The broadband IC amplifier is used as an rf amplifier.

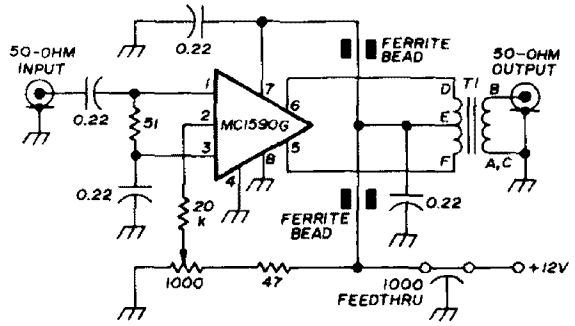


fig. 1. Broadband IC amplifier covers the frequency range from 1 to 56 MHz and provides approximately 25-dB gain with a noise figure of about 13 dB at 30 MHz. Transformer T1 is a broadband 400-ohms balanced to 50-ohms unbalanced type such as the North-Hills 556026 available from the author. All 0.22-μF capacitors are Erie Redcaps (see text).

With such a temporary setup you *may* be able to take a first cut at reception of the satellite. If you used 47 MHz as the LO frequency the image at 54 MHz (another ham operator) might provide some occasional unwanted interference,

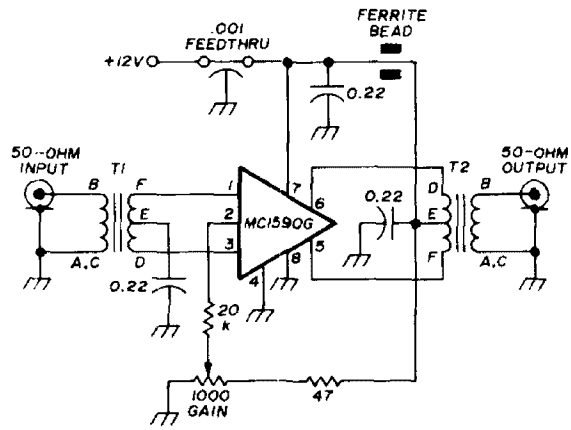


fig. 2. This broadband IC amplifier uses two broadband transformers to provide 40-dB gain with a noise figure of about 3 dB at 30 MHz. Transformers T1 and T2 are broadband 400-ohms balanced to 50-ohms unbalanced types such as the North-Hills 556026-336 available from the author. All 0.22-μF capacitors are Erie Redcaps (see text).

but at least you would have an operating receiver system.

A second illustration is one where a measurement problem exists. A transistorized heterodyne exciter is being built. In the exciter two oscillators are mixed together and they are *supposed* to provide an output at 29 MHz from the mixer. However, it is possible to have mixed a harmonic from one of the oscillators (or both), and it is also possible to get either the sum or the difference frequency out of the mixer. You want to use a frequency counter on the output of the mixer. However, the mixer output level is far too small to operate the counter. By putting the broadband IC amplifier after the mixer, the output can be brought up to 0 dBm (0.225 volt rms across 50 ohms). This will operate most counters.

The first illustration is a case where it would be desirable to use the more

complicated broadband amplifier because of its lower noise figure. In the second example the more simple broadband amplifier probably would have been adequate. The single-transformer amplifier, having a 51-ohm resistor across its input terminals, represents an almost pure resistive impedance that is quite unlikely to upset the circuit being tested (if it is made to drive 50 ohms).

It was not intended in either example that the use of broadband amplifiers

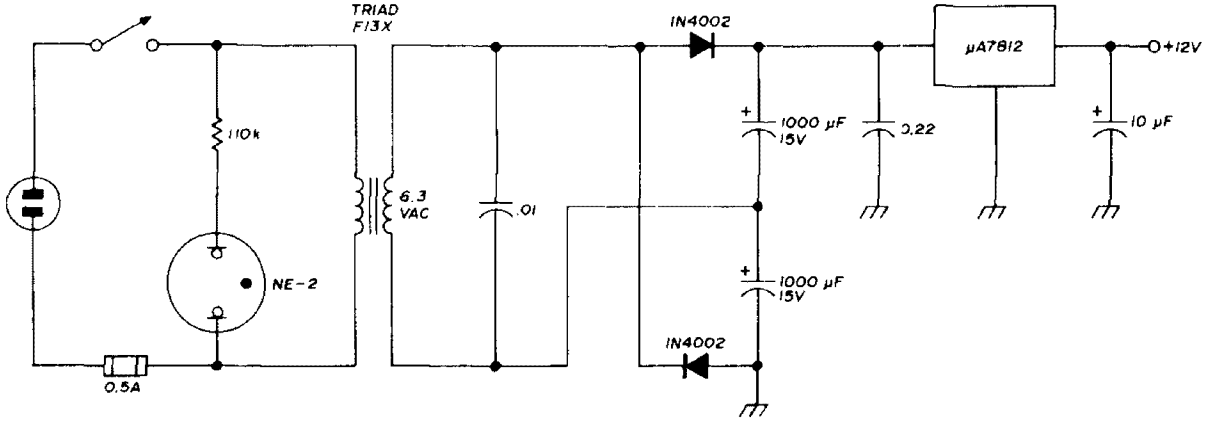


fig. 3. Simple regulated power supply for the broadband amplifiers uses a Fairchild μA7812 voltage-regulator IC (Fairchild part number UGH7812393).

represent an optimum solution, but rather, a quick one – from which an assessment of system problems can be made.

Whether or not a broadband amplifier should be used as a preamplifier ahead of a communications receiver is open to question. It is often true that you get a better noise figure with a preamp ahead of the receiver and therefore, better sensitivity. However, the extra gain of the preamp will, in general, reduce the dynamic range of the receiver. This usually has the consequence of causing cross-modulation in some stage of the receiver.

construction

Both broadband amplifiers are built into cast-aluminum Pomona 2905 boxes which measure about 1½x2½x4¼ inches. These boxes are normally furnished with an aluminum top plate. In these amplifiers this plate is replaced by a piece of double-sided copper printed circuit board. The copper laminate makes an excellent ground plane and provides good grounding. Note that pins 4 and 8 of the IC socket are bent down and soldered to the copper ground sheet. This provides both a low inductance ground and mechanical support for the socket.

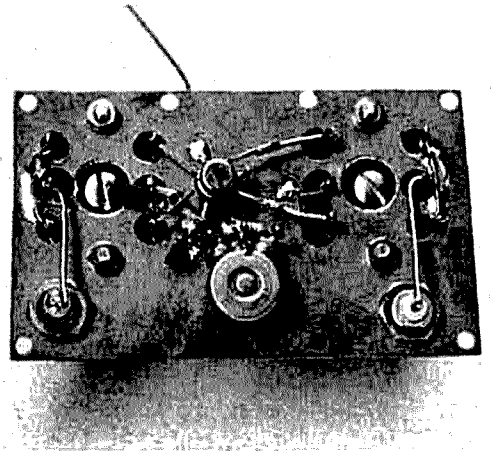
The sockets I used for the MC1590G ICs are Robinson-Nugent type LP-5178. The 400-ohm balanced to 50-ohm unbalanced transformer is similar to the North-Hills 0600AA. The 400:50-ohm transformer I used is a special unit (556026-336) that I obtained as a project overrun item.* The transformers are mounted above the copper-laminate plate with their pins protruding through to the circuitry side. Pins A and C are grounded and pin B connected to the adjacent BNC jack. Pin E is the center-tap of the balanced 400-ohm side with D and F the high ends.

The long spider-like legs of the IC socket (pins 5 and 6) are bent a bit and soldered directly to the D and F pins of the output transformer. This provides short, low-inductance connections and gives additional support to the IC socket.

Since the amplifiers must operate over wide bandwidths, it is important to use

*Available for $5.00 plus 50 cents postage, each, from Hank Olson, Box 339, Menlo Park, California 94025.

relatively large-value, low-inductance capacitors for rf bypassing and coupling. The particular types of 0.22-μF capacitors I used are Erie Redcaps which are small ceramic units. If these capacitors cannot be readily obtained it would be better to use 0.01-μF or 0.02-μF ceramic discs that to use a standard foil-type 0.22-μF capacitor such as a Mylar type. The smaller 0.01- or 0.02-μF disc ceramic capacitors may cause some loss of gain at the lower end of the passband, but that's



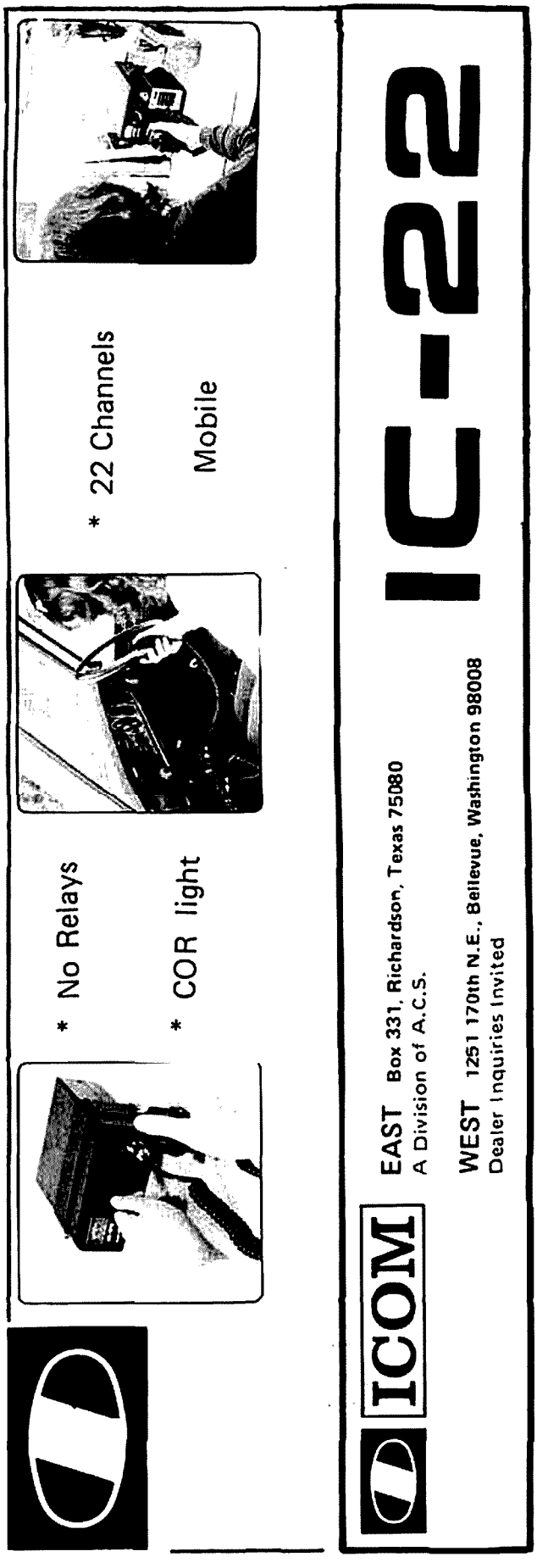
Construction of the broadband IC amplifier. Note the simple, direct-path wiring for minimum inductance.

preferable to having highly inductive capacitors at the high frequency end.

The ferrite beads used to decouple the +12-volt supply line between different bypassed points in the circuit are made by a number of manufacturers, including Ferroxcube, Stackpole and Allen Bradley. They are also available in smaller quantities from Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607.

A simple regulated power supply for the IC amplifiers is shown in fig. 3. Although the supply is capable of considerably more output current, the amplifiers require only about 40 mA. The use of a 6.3-volt ac filament transformer, a full-wave voltage doubler and an inexpensive three-terminal voltage-regulator IC make this supply not only a good performer, but also low in cost.

ham radio



using the Heath SB-650 frequency display with other receivers

Complete details on modifying the popular SB-650 for use with the Allied SX-190 and other receivers

Marvin R. Clinch, K2BYM, 4500 Skenandoah Drive, RD 2, Oneida, New York 13421

The introduction of the Heath SB-65(frequency display unit has provided ; relatively easy means of adding a digita frequency display to the Heath SB anc HW receivers and transceivers. The SB-650 provides a decimal readout of the frequency to which the receiver or tran sceiver is tuned to the nearest 100 Hz This article describes a means of using the display with virtually any receiver with out permanent modification of the SB-650. The entire modification effor takes less than an hour and can be removed in five minutes if you wish tc return the frequency counter to its origi nal configuration.

the counter

The SB-650 is designed to accept three input signals from each of the three oscillators present in Heath receivers anc transceivers. These are the high frequency oscillator (hfo), a switched fixed fre quency oscillator; the linear master oscil lator (LMO), a variable frequency oscilla tor of limited range; and the bea frequency oscillator (bfo), an oscillato either fixed or variable, usually arounc 455 kHz.

The heart of the SB-650 frequency display unit is an integrated-circui up/down counter which provides the capability to count and add or subtract frequencies. The Heath design counts the hfo frequency, then counts down tc subtract the LMO frequency. The resul of this set of operations is then displayec on a set of Nixie digital displays. The

Nixie tubes display the operating frequency to the nearest 100 Hz.

mixing combinations

A block diagram of a hypothetical receiver with only the oscillators and mixers is shown in fig. 1. There are a number of ways the oscillator frequencies may be set to receive a given frequency signal and still produce the same frequency output. First, let us assume that the last i-f is 455 kHz. The bfo frequency is also 455 kHz, plus or minus a few kHz. For the purposes of discussion assume that it is exactly 455 kHz, thus providing a zero beat with any incoming signal which is precisely tuned to the center of the i-f bandpass filters. More on this later.

Now, the mixing of two signals to generate a desired frequency can be shown in eqs. 1 and 2.

$$F_3 = F_2 - LMO \quad (1)$$

$$F_3 = LMO - F_2 \quad (2)$$

In these equations, only the difference frequency is considered. It is assumed that appropriate filtering is available to remove the sum and fundamental frequencies. Thus, F_3 (455 kHz) can be generated if the LMO frequency is 455 kHz below the input signal frequency (eq. 1) or if the LMO is 455 kHz above the signal frequency (eq. 2).

Similarly, F_2 can be generated if the hfo frequency is either above or below the incoming frequency (F_1) by F_2 (eqs. 3 and 4

$$F_2 = F_1 - HFO \quad (3)$$

$$F_2 = HFO - F_1 \quad (4)$$

By substituting and re-arranging terms, it is possible to come up with four different methods of finding the input frequency F_1

$$F_1 = HFO + BFO + LMO \quad (5)$$

$$F_1 = HFO - BFO + LMO \quad (6)$$

$$F_1 = HFO - BFO - LMO \quad (7)$$

$$F_1 = HFO + BFO - LMO \quad (8)$$

$$F_1 = HFO - BFO \quad (9)$$

$$F_1 = HFO + BFO + LMO \quad (10)$$

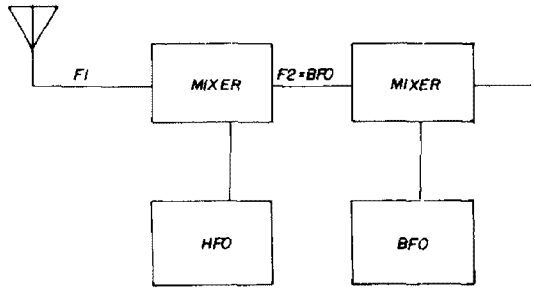


fig. 2. Simple single-conversion receiver uses only two local oscillators.

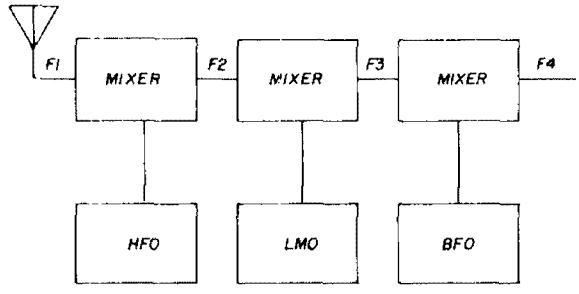


fig. 1. Simplified block diagram of a double-conversion communications receiver. Signal frequencies F1, F2, F3 and F4 are discussed in the text.

If you look at these equations for a minute you'll see that the frequency to which the receiver is tuned can be found by simply counting and adding or subtracting the various oscillator frequencies. This in no way requires an input signal from the first rf stage of the receiver. The Heath SB-650 is designed to operate with equipment meeting the parameters of eq. 7.

modifications

With only slight modification of the input and control circuits it is possible to use the SB-650 with receivers configured to meet eqs. 6, 8 or 10. A single-conversion receiver meeting eq. 9 and configured as shown in fig. 2. may be accommodated without modifying the SB-650.

The counter portion of the SB-650 makes a complete count every 160 milliseconds. This time is divided up into four equal periods of 40 milliseconds each, t_1 through t_4. During the first three of these periods the counter counts the appropri-

ate frequencies, and during the last, the final count is transferred to the display. A pair of control signals are generated by the digital timing circuits every 40 milliseconds to control these operations.

Each IC counter (SN74192N) has two input lines. With the proper manipulation of these input lines, the counter can be made to count up, count down (by

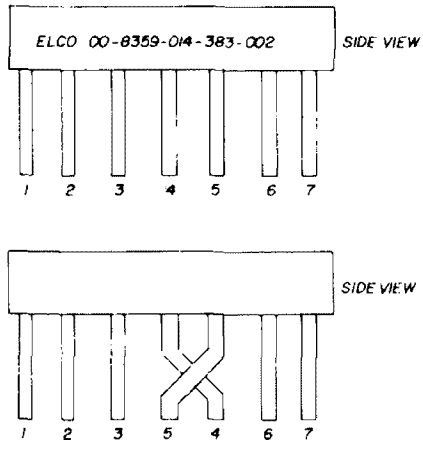


fig. 3. Modification of a 14-pin IC socket to change the counting arrangement of the SB-650. IC-12 is removed from the SB-650, this midified socket is plugged in, and IC-12 is plugged into it.

counting backwards), or not to count at all, as shown in table 1. Since the control signal is digital, the digital convention is used, a 1 being a high voltage (5 volts nominal for the SN7400 series of ICs) and a 0 being a low voltage (zero volts nominal). These logic signals are applied to the counter inputs to cause the counters to count up, down or to inhibit the count. The *signal* of table 1 consists of the frequency being counted. This signal is previously shaped to consist of alternate ones and zeroes.

The SB-650 has one period of count up and two periods of count down. For the Allied SX-190 receiver you need *two* periods of count up and *one* period of count down. To make the change simply follow the instructions. Take one 14-pin IC socket (ELCO 8359, 014-383-002) and bend the socket pins 4 and 5 as shown in fig. 3B. Put a piece of insulated tubing over one of the two leads just bent (heat-shrink tubing is best). File each lead to a taper so that it may be inserted into IC socket in the SB-650. Now, remove IC-12 from the SB-650. Plug the modified socket into the IC-12 socket and plug IC-12 into the modified socket. The counter will now count up for two periods and down for one period. There is no change in the display cycle.

After the modification, the hfo input signal counts down while the bfo and LMO inputs are counted up. This does not yet meet the requirements for the SX-190. However, this is no real problem. You only have to shift the input circuits around. The hfo input has a selected fet for better high-frequency response, so you want to retain it as the hfo input (count up). To do this, you have to make two more jumpers for the appropriate logic inputs and outputs.

Take two more 14-pin IC sockets and bend the pins outward as follows: on one socket bend pins 2 and 6 as shown in fig. 4. On the other socket bend pins 3 and 13 as shown in fig. 5. File the remaining socket pins so they can be plugged into another IC socket. Now solder a 4- or 5-inch piece of insulated hookup wire to each of the bent pins on the first socket. Cover the soldering pins with insulating tubing. Tin the other end of each wire.

Slip another piece of spaghetti over the wires and solder the wire from pin 2 of the first socket to pin 3 of the other socket. Solder the wire from pin 6 of the first socket to pin 13 of the other socket. Slide the spaghetti over these connections. Remove IC-24 from the SB-650 and insert the first modified socket in its

table 1. Counting periods and inputs to the Heath SB-650 frequency display unit.

	time	input 1 (pin 4)	input 2 (pin 5)
T1	count up	1	signal
T2	count down	signal	1
T3	count down	signal	1
T4	display	1	1

place. Insert IC-24 into the new socket. Now remove IC-10 and insert the second modified socket in its place. Re-insert the IC-10 in its new socket. Now connect the terminals on the back of the SB-650 as

shown in table 2. With the counter modified in this way and the input connections shown in table 2 you can use the counter for receivers which are designed to meet eqs. 6 or 8. The Allied SX-190 follows eq. 8. Simply connect the receiver outputs to their respective counter input terminals and use it as described in the SB-650 manual.

receiver oscillators

The SB-650 input uses a high impedance fet. The Heath Company recommends you use 50-ohm coaxial cable to connect the receiver to the counter (RG-58A/U). It is possible that the loading effects of the coaxial cable may stop the oscillator from oscillating. Such appeared to be the case with the SX-190 bfo even though the counter was not connected to the cable. The buffer amplifier shown in fig. 6 was connected to the bfo through a 7-pF capacitor to reduce any loading on the bfo. With this buffer in the circuit the oscillator worked without any problems. The same fet buffer circuit may be used for other oscillators which are sensitive to loading.

tuning

Since the SB-650 was designed primarily for ssb equipment with fixed beat-frequency oscillators, proper technique must be used when the receiver has only a variable bfo, and is used on CW or a-m signals. First, tune the receiver to the desired signal for maximum signal strength (that is, tune it to the center of the receiver i-f passband). Now, turn the bfo on and zero beat it with the signal. Now the incoming signal (after conversion) and the bfo are exactly the same frequency, and the SB-650 will be able to read the incoming frequency within the nearest 100 Hz. (The human ear can zero beat well below 100 Hz difference, so you don't have to worry about accuracy.)

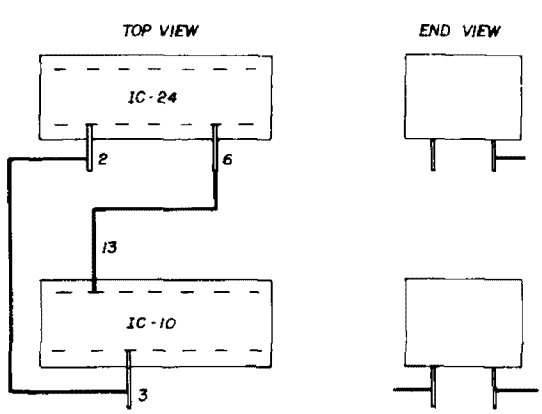


fig. 4. Simple technique for rearranging the input circuitry of the SB-650 (see text).

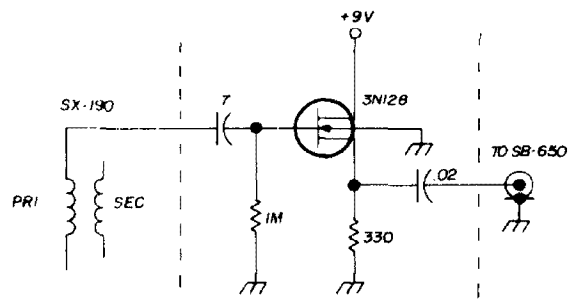


fig. 5. Buffer amplifier may be useful where coaxial cable loads down the oscillator circuit.

If you are receiving an a-m signal turn off the bfo and note that the frequency counter has changed frequency by the amount of bfo frequency. If you are receiving a CW signal, when you tune the bfo to provide a good CW tone the counter display will change by the bfo frequency change. Just remember that for a proper count the variable bfo must be turned on and zero beat with the incoming signal.

other receivers

The SB-650 frequency counter can also be used for single-conversion receivers which meet either eq. 9 or 10. It should be cautioned, however, that for a band-switching receiver all of the bands must meet the same set of mixing relationships.

For a receiver meeting eq. 9, simply connect the hfo and bfo from the receiver to the counter. It would also be best to terminate the LMO input with a 50- or 52-ohm resistor to prevent noise spikes from causing faulty readout. If you have a single-conversion receiver meeting eq. 10 modify the counter as described above and connect as shown in table 2 for eq. 10.

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your receiver is designed? Just use the counter to measure your local-oscillator frequencies. Terminate the LMO and bfo inputs with 50-ohm resistors. Now, tune the receiver to some appropriate frequency and connect the counter hfo input to each of the receiver oscillators. Calculations will show the frequency arrangement of the receiver and which equation the receiver meets.

other uses

The Heath SB-650 frequency display is a most versatile piece of equipment and can be used with many receivers. It can also be used as a standard frequency counter up to 30 MHz with resolution of 100 Hz. One unit tested by WA2DJU ran up to nearly 50 MHz with only a slight degradation of sensitivity. Transmitter frequencies may also be counted if the signal is kept to a point where the counter is not damaged. However, take precautions to make sure that the ratings quoted by Heath are not exceeded. The input mosfet is tough but it cannot be abused.

table 2. Input connections to the modified SB-650 frequency display unit.

SB-650 input	eq.6	eq. 8	eq. 10
HFO	HFO	HFO	none
BFO	LMO	BFO	BFO
LMO	BFO	LMO	HFO

Input connections to unmodified SB-650

SB-650 input	eq. 7	eq. 9
HFO	HFO	HFO
BFO	BFO	BFO
LMO	LMO	none

I would like to express my appreciation to Edward A. Murphy, a serious SWL, who brought up the original question about the SB-650's added capabilities and who was willing to gamble on my suggestions. Thanks are also due to Mr. E.J. Cupples, WA2DJU, who tested the frequency response of the SB-650 and provided the fet buffer amplifier circuit.

ham radio

logic oscillator

for multi-channel crystal control on vhf fm

Using digital TTL ICs in a logic oscillator to provide a bank of selectable crystal frequencies

Stirling Olberg, W1SNN, 19 Loretta Road, Waltham, Massachusetts 02154

The acquisition of several Motorola fm mobile rigs, which are single-channel vacuum-tube rigs, required modification to include at least six additional receiving and transmitting frequency controls. Since the space to include the additional tubes and components is at a premium within these chassis, TTL ICs seemed to be appropriate for the additional circuitry.

These components are mounted on a small board which has a modified FT-243 crystal holder, epoxied to one end, allowing it to plug into the Motorola crystal socket. The wiring to the 12-volt power supply and a frequency selector switch have separate connections which will be described later.

the circuit

The circuit consists of six individual crystal oscillators which may be adjusted exactly to the required frequency by a trimmer capacitor, C1. The input to each oscillator is turned on when the selector switch is set to the desired crystal frequency. Only one of the oscillators is in operation at one time, as will be explained using channel 1 as an example. In this discussion a logic 1 is a plus voltage and a logical zero is ground.

The frequency selector switch is set to channel 1, grounding the logical 1, provided by R13 at the input of inverter U5, to a logical zero. The conversion at the output of U5 presents a logical 1 to an input of U1, turning on the oscillator. The oscillator output is fed to another inverter, U4, which presents a buffered output to the receiver or transmitter it is used with. All channel outputs are in parallel.

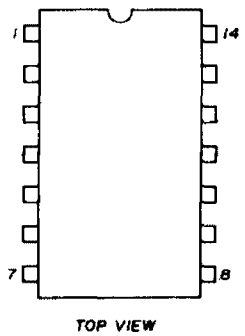
A TTL logic level of 3 volts average is the rf level at the combined oscillator outputs; this provides sufficient drive to the original oscillator grid. This output is a square wave but offers no problem, because the original vacuum-tube oscil-

lator now acts as a buffer and since it has a tuned output circuit, it provides a sinusoidal output.

Five-volts dc is required for the logic circuitry. This is controlled by the simple voltage regulator built around transistor Q1. This regulator insures that the 5 volts for the logic circuits will be held stable when operating from an auto-battery charging system.

installation

The control wiring in my car consisted of a section of seven-conductor shielded intercom cable. The frequency selector switch is located at the control head, near the front seat of the car. In this installation the receiver and transmitter have independent frequency controls allowing cross channel listening and operating. If this feature is not desirable, the two switches may be ganged together. The 12-volt dc power was supplied from the filament string.



summary

It might seem, at first glance, that this offering does not add much to improve a piece of equipment simply because it uses ICs and some well known logic facts. The fact is, however, that it adds an indefinite number of channels, in a very limited space, with a modest amount of effort. It can be built and installed in one evening's work into almost any receiver or transmitter which requires additional channels.

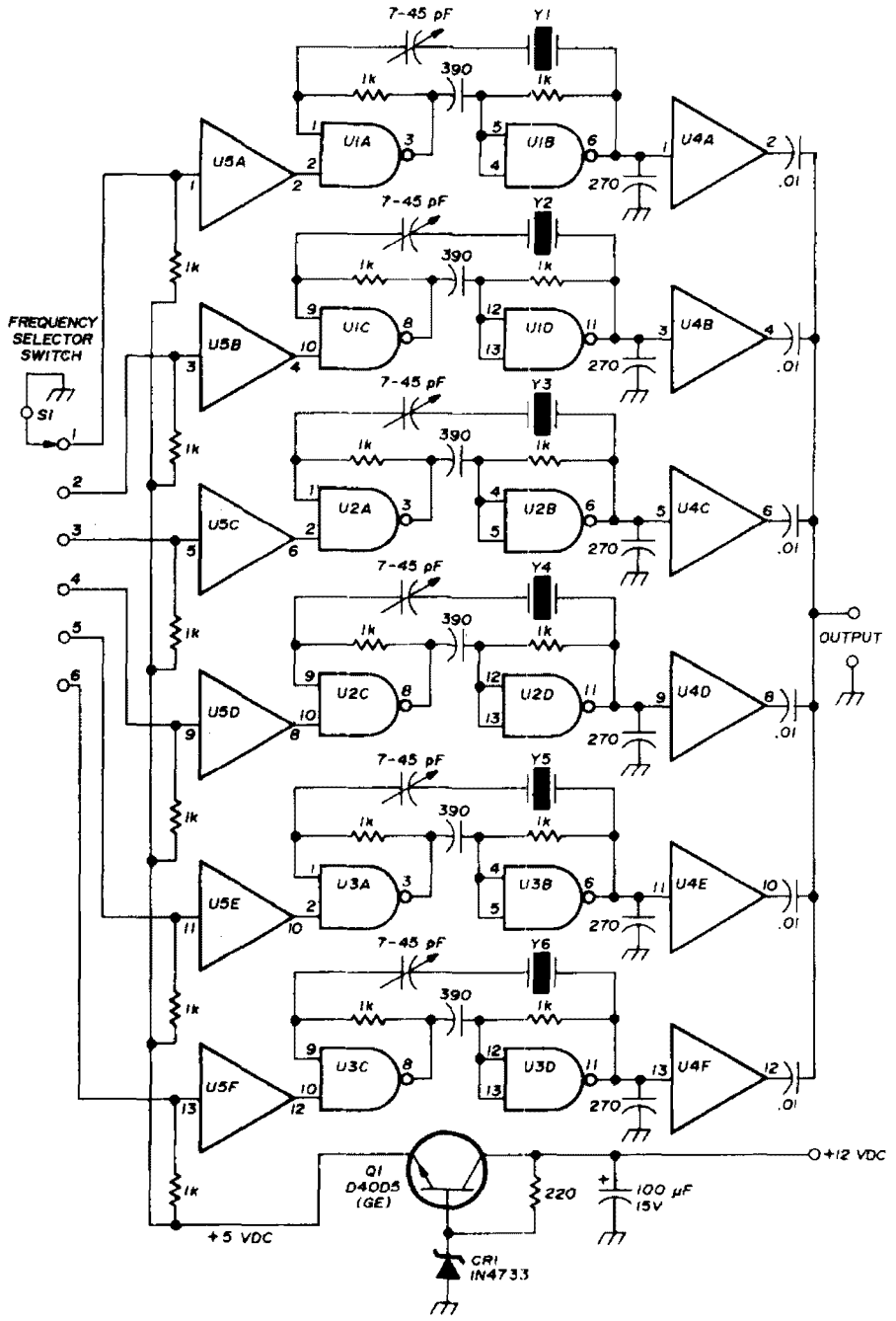
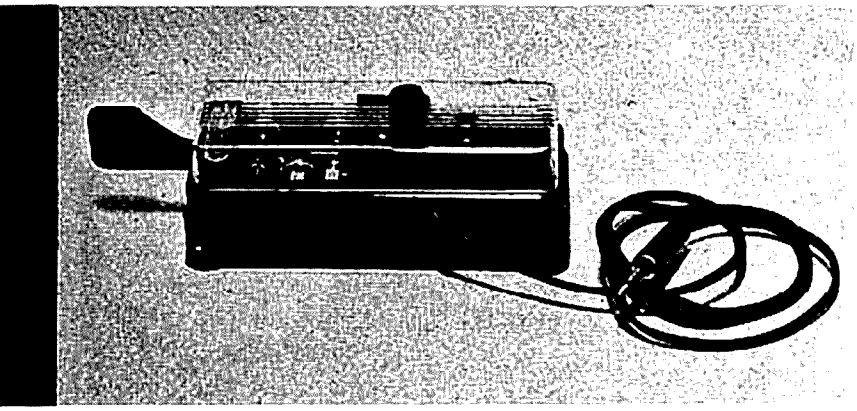


fig. 1. Schematic diagram of the 6-channel crystal-controlled oscillator. Simple control is possible because of the logic of the TTL ICs which are used in the circuit. ICs U1, U2 and U3 are SN7400N quad NAND gates; U4 and U5 are SN7404 inverters. Crystals are Jan Crystals type 4C-6/U, frequency as required. Switch S1 is a Mallory 8511496. Although not shown on the drawing, pin 14 of all ICs is connected to the regulated 5-volt line; pin 7 of all ICs is connected to ground.

ham radio



solid-state bug

An improved version of the semi-automatic key using the principles of the solid-state switch

Arthur J. Glazar, K2FV, 31 Amapola Lane, Kings Park, New York 11754

Automatic CW generators are not for me. I enjoy hand sending and suspect there are many other operators who are turned off by anything more automated than a bug. But the bug can stand improvement, and this article describes one operator's approach to modernizing the semi-automatic key. I wanted to maintain the bug's basic operating characteristics and features (automatic dots, manual dashes, and portability) while eliminating its shortcomings (mechanical dot system, unreliable contacts, and awkward speed control). The design goals were: conveniently adjustable speed during operation, all solid-state design, low-drain battery operation, single unit construction and a "feel" identical to a conventional bug. The result is the solid-state bug shown in the photograph. The circuit is built into a modified Lafayette bug.* The dot spring and damper were removed to make room for the five-transistor circuit shown in fig. 1.

features

Dot speed is knob-controlled and adjustable from 4 to 22 dots per second. The dot weight control is a screwdriver adjustment to avoid confusion during operation and is variable from about 25 to 100 percent weight. Once set, the

*Lafayette Radio Electronics Catalog number 99P25520, $14.75.

weight remains constant over the speed range. At maximum weight the dots merge, and the key becomes a sideswiper rather than a bug. This mode of operation, may hold some appeal to mobile marine operators.

The circuit operates from a 9-volt transistor radio battery and draws no current during key-up conditions, thus assuring long battery life and eliminating the need for an on/off switch. The output

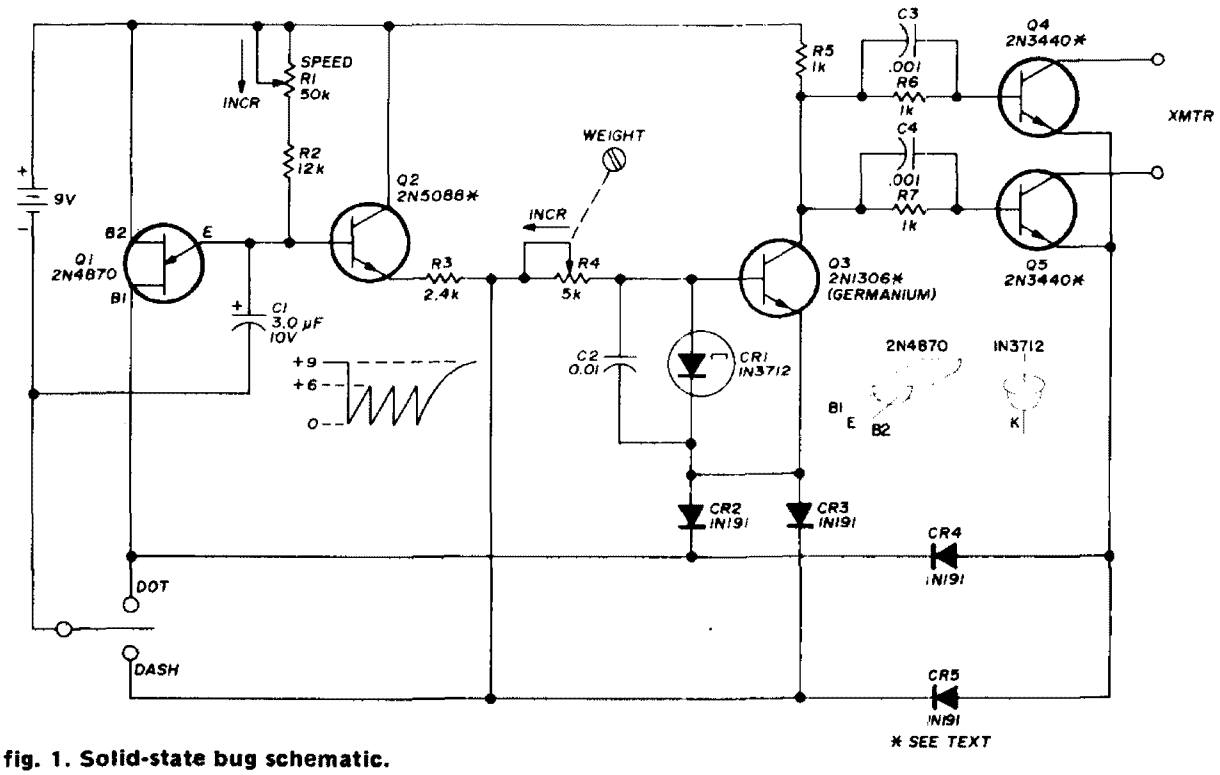


fig. 1. Solid-state bug schematic.

of the bug is, in effect, that of a solid-state relay. It is a bilateral transistor switch capable of keying circuits up to 250 volts at 15 or more mA. Polarity need not be observed when connecting the bug to the keyed circuit. Transistors less expensive than the 2N3440 may be substituted for Q4 and Q5 if only low-voltage circuits are to be keyed. The 2N3440 has a V_{CBO} of 300 volts.

The amount of current that can be keyed is determined by resistors R5, R6, and R7. With the values indicated (1k) the circuit should handle at least 15 mA and perhaps double that amount; the maximum being a function of transistor gain.* To double the current-carrying capability the resistor values should be halved; to halve the current the resistor values should be doubled, and so forth. Because battery drain increases in direct proportion to current requirements, the largest resistor values should be used. With the values shown, key-down battery drain is less than 6 mA. At that rate, battery life should approach shelf life.

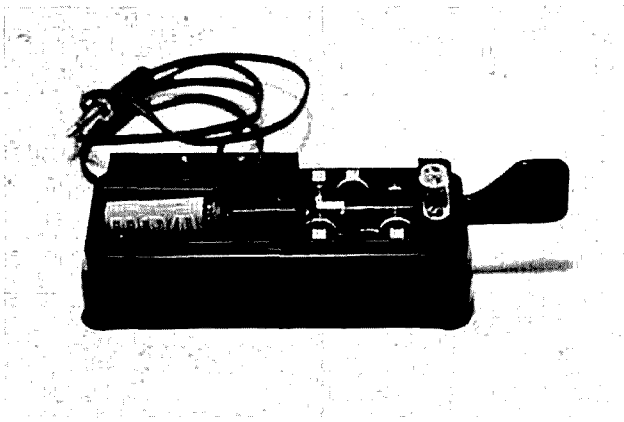
Note that bilateral operation of the key depends on the ability of the battery to float with respect to the keyed circuit. Therefore there must be no direct connection between the battery and any transmitter voltage bus, including ground.

circuit

The bilateral output swtich[1] is closed when Q3 is off, and base current is therefore supplied to Q4 and Q5 via R5, R6, and R7. When a dash is made, Q3 is turned off by returning the junction of R3 and R4 to battery negative, thus removing its base drive. Closing the dot contact starts the dot generator, Q1,

*Actually, the *inverted* gain of Q4 and Q5. See reference 1.

which applies negative-going dot pulses (via emitter-follower Q2) to the base of Q3, thus turning it off. The dot generator is a unijunction transistor relaxation oscillator.[2] It is idle until the dot contact is closed, which connects the base 1 (B1) terminal to battery negative and permits it to oscillate. As long as the dot contact is closed, the voltage on C1 approximates a linear saw-tooth waveform, oscillating between about 0.6 volt (with respect to



The bug with plastic cover removed, showing the battery at left front, the speed-pot shaft protruding vertically from the chassis, and the weight control to the left of the speed pot.

battery negative) and 6.2 volts, the exact amplitude being a function of Q1 parameters.

The oscillation frequency is determined by R2, C1, and the speed-control pot, R1. The sawtooth on C1 is applied through emitter follower Q2 to a variable threshold level detector comprising R3, R4, tunnel diode CR1, and transistor Q3.[3] The high input impedance of the emitter follower (roughly 700k) prevents capacitor loading.* The voltage at which the level detector trips is determined by the setting of the weight control, R4. Q3 is turned off when the sawtooth is below the preset level. The higher the threshold level (i.e., the higher the value of R4), the longer the dot duration becomes.

*The 2N5088 is an inexpensive plastic transistor with an h_{FE} of about 300. Any other npn with a gain approaching 100 can be used for Q2. Q3, however, must be a *germanium* npn.

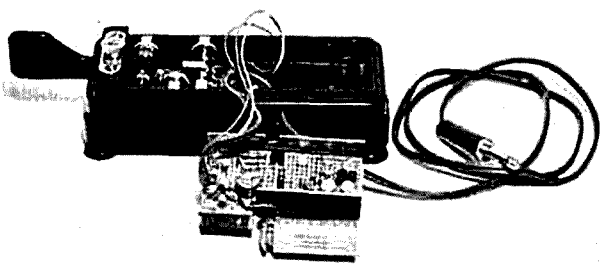
Note that Q3 emitter is permitted to return to battery negative only during keydown: through CR2 for dots, and through CR3 for dashes. Thus, Q3 cannot draw current during keyup. A similar situation exists for the emitters of Q4 and Q5, which return to battery negative through CR4 for dots and through CR5 for dashes. The purpose of this latter arrangement is to *prevent* the dots from self-completing, thereby preserving the feel of a standard bug.

Capacitor C2 is essential to prevent stray rf from tripping the tunnel diode. Capacitors C3 and C4 were found, by trial and error, to be most effective in further desensitizing the circuit to stray rf.

construction

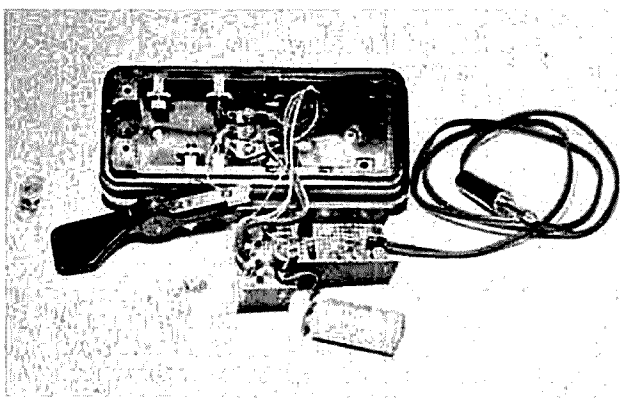
The circuit is built on a small L-shaped piece of 0.1-inch perforated board. To keep component height to a minimum no sockets are used. The board is mounted upside-down on 5/16-inch spacers inside a U-shaped chassis. The inverted U rests inside the bug's base and conceals all the circuitry. An extra fold in the U permits easy access to the battery. The chassis simply rests in the base with no fastening; the plastic cover affords adequate hold-down, since it fits snugly on the base. For traveling, a rubber band around the entire unit works nicely to keep everything secure.

The speed-pot shaft extends through the plastic cover, and the weight adjustment simply requires a hole for screwdriver access.



Circuit components shown when the chassis is lifted from the base and inverted.

The photos should give sufficient information to duplicate the solid-state bug. One suggestion may save time for someone attempting to disassemble a Lafayette bug; the base is weighted with three pieces of sheet iron, which fit snugly in the underside of the base. They are held by two screws and considerable friction. (The screw heads are visible in the top view photo of the disassembled bug.) Once the four rubber feet and the



A top view of the disassembled bug. Two Phillips-head screws can be seen, which hold the sheet-iron weights in the base. Toward the right can be seen slots where original binding posts were removed to make room for the chassis.

two screws are removed, the sheet iron must be forcibly dislodged by poking through the screw holes and by slamming down the base.

With the sheet iron out of the way, the contact wiring is accessible and must be modified to spdt operation. I found it necessary to carve away some of the plastic base material to accommodate the re-routed wiring, but this was easily accomplished with a sharp wood chisel. The dot, dash, and common connections are brought to the top surface of the base by means of three countersunk flathead 6-32 screws. On top, solder lugs are used to make connection to the circuit.

references

1. "GE Transistor Manual," (Seventh Edition) General Electric Company, Syracuse, New York, 1964, Chapter 3.
2. Ibid., Chapter 13.
3. Ibid., Chapter 14.

ham radio

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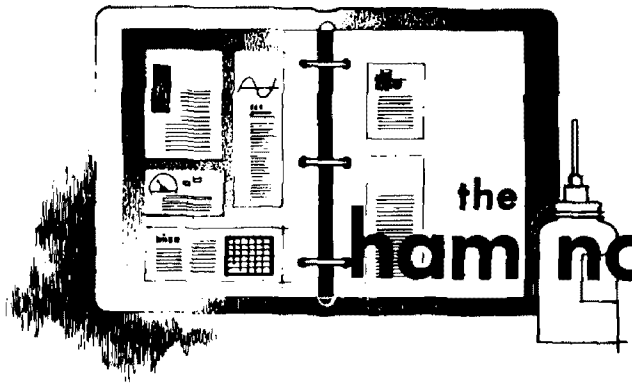
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crystal control and narrow shift RTTY with the Heath SB-series

The RTTY enthusiast interested in narrow shift auto-start faces two problems with the Heath SB-Series of receivers and transceivers. The first problem is modifying the receiver to receive narrow shift with the optional 400-Hz filter. This filter, unlike the ssb filter, is centered at 3395.4 kHz and will not pass the standard tones of 2125 and 2295 Hz. It is therefore necessary to move the bfo injection frequency to 3393.190 kHz. This frequency is somewhat critical and dependent on the actual filter bandpass. It is possible to use a bfo crystal provided with the receiver to provide the proper injection frequency.[1]

Although the Heath LMO is known for its extremely good stability, it will not hold the frequency well enough to provide unattended auto-start operation over a period of weeks or months. There are two solutions to this problem. One is to crystal lock the receiver. This is not quite the same thing as crystal control. In the crystal lock mode, a crystal is tied from the grid to ground of the variable oscillator tube. The oscillator tuning then allows varying the frequency about 50 Hz either side of the frequency determined by the crystal. The fine tuning effect and the simplicity of this method make it quite desirable.

On the negative side, the crystal lock method does not allow for quick change from one auto-start frequency to another, and in many cases it is not possible to pull the crystal far enough to hit the desired frequency. It is usually possible to pull the frequency about 50 Hz. It may be difficult to purchase a crystal this close to the desired frequency. Also, crystals and oscillators have been known to change enough that the crystal lock method fails after a time. However, this is unusual.

I had the problem that the high-quality (high-price) crystal I ordered would not hit the frequency in the crystal locked mode. I decided that the XT-4 oscillator[2] with its low-capacitance circuit would allow more variation in output from a given crystal than any other oscillator that I was familiar with.

The XT-4 was built, and the output frequency was adjusted to center the receiver on the auto-start frequency. I had the receiver crystal controlled on frequency now, but receiver sensitivity was degraded by insufficient mixer injection from the XT-4. Several cures were tried, such as tuned output on the XT-4, but injection was insufficient in all cases until a buffer stage was added.

Fig. 1 shows the modified XT-4 circuit with the frequency-shift keyer removed. It would be advisable to review Hoff's article on the XT-4 concerning parts and

calculation of crystal frequency. He also gives ordering information for the crystals which should be followed carefully. For general information, I ordered a 5425.0 crystal for the 14075 kHz auto-start frequency. When placed in the XT-4, and properly centered on frequency, the crystal oscillator output measured 5424.229 kHz. It was not possible to pull the crystal this far (almost 800 Hz) in the crystal lock mode, but was no problem at all with the XT-4. In fact, I could have moved it considerably farther.

The bandpass transformer, T1 is available from Heath (No. 52-103) for $4.30. The rf output from the modified XT-4 is routed to a ceramic switch which selects either the original LMO or one of several XT-4 oscillators. The oscillator, buffer, power supply and regulator were built in a 4 x 6 x 3-inch Minibox. It is a bit crowded, but everything was made to fit. It is important that the crystal be shielded from rf. If the oscillator is used in a room where wide temperature variations occur, it may be necessary to temperature compensate the oscillator to realize the high degree of stability necessary for unattended auto-start.

Robert Clark, WA4VYL

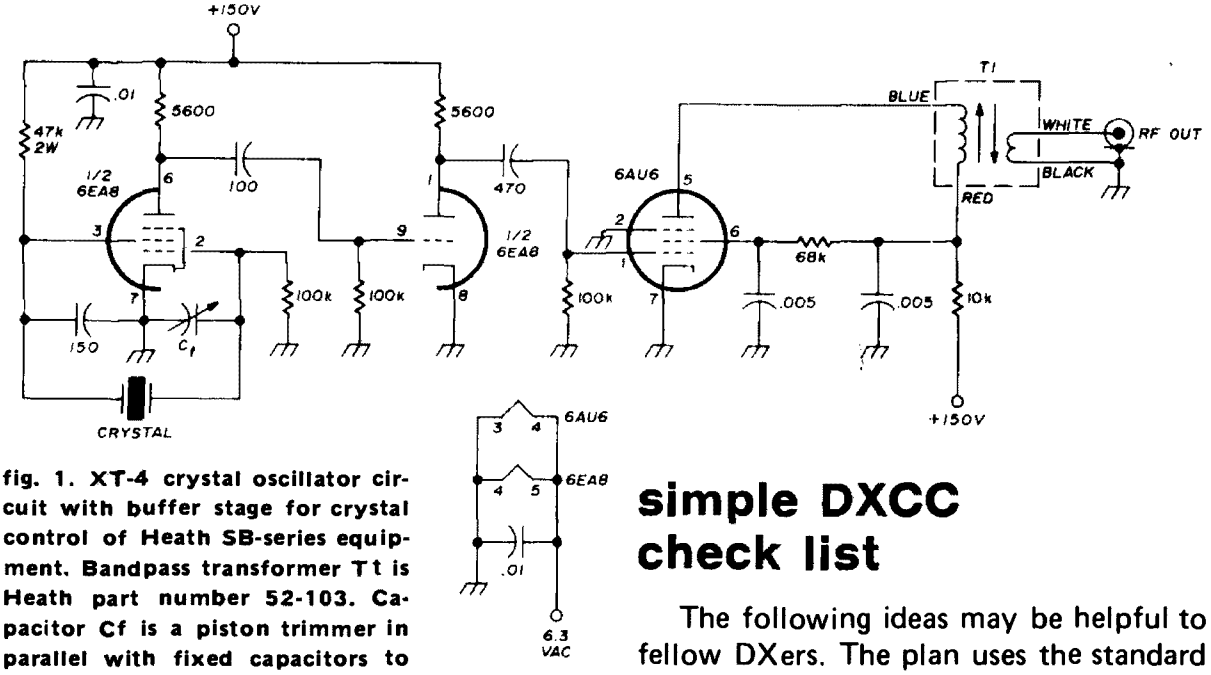


fig. 1. XT-4 crystal oscillator circuit with buffer stage for crystal control of Heath SB-series equipment. Bandpass transformer T1 is Heath part number 52-103. Capacitor Cf is a piston trimmer in parallel with fixed capacitors to adjust oscillator to desired frequency.

references

1. Robert Clark, K9HVW, "Narrow-Shift RTTY Reception with Heath SB Receivers," *ham radio,* October, 1971, page 64.
2. Irvin M. Hoff, W6FFC, "Crystal Controlled RTTY," *RTTY Journal,* December, 1967, page 4.

simple DXCC check list

The following ideas may be helpful to fellow DXers. The plan uses the standard ARRL Countries List (Operating Aid No. 7). By simply marking it appropriately, immediately to the *left* of the call-letter prefix which is followed by the country identity, you have a progressive and updated record of your DXCC status. You know at a glance if the country is: needed, worked, confirmed (QSL received), or credited (officially counted).

The legend of symbols I use for this is as follows:

- = worked
+ = confirmed
⊕ = credited to DXCC count

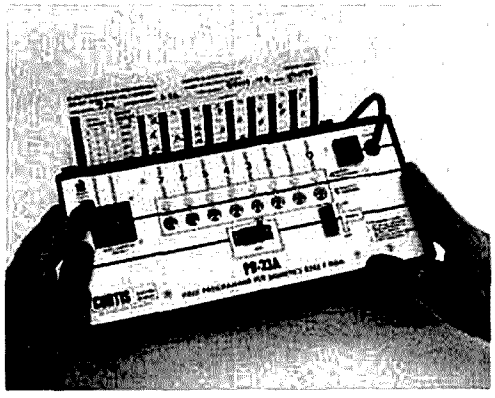
To re-cap, a short horizontal mark for *worked,* add a short vertical mark through the horizontal mark for *confirmed* when the QSL card is received, and circle the ⊕ when official credit is received.

As you near Honor Roll status, you will want to identify those countries which have been deleted as current countries. Simply draw a slightly larger circle to form a double-circle symbol to easily spot those deleted countries.

Lowell White, W2CNQ

new products

ROM programmer



An inexpensive portable field programmer for the Signetics 8223 256-bit field-programmable ROM has been introduced by Curtis Electro Devices. Called the PR-23A, this instrument allows manual programming of the 8223 in the laboratory, in the factory or in the field by unskilled personnel.

To program a device, the operator places a blank F-ROM in the test socket, selects the appropriate octal word address and presses one of eight output push-buttons to open the selected fusible link. Program sequence is automatic and independent of operator timing to ensure uniform program conditions. A typical pattern can be programmed in about five minutes.

Eight indicator lamps provide a continuous display of output states. A slot at the rear of the instrument holds the truth table in proper alignment with panel markings.

The PR-23A is priced at $199.50. A 240 Vac model also is available. For more information contact Curtis Electro Devices, Inc., Box 4090, Mountain View, California 94040, or use *check-off* on page 94.

rackmounted directional rf wattmeter

The new Bird model 4372 Thruline® Directional High Power Wattmeter covers 25 to 500 MHz from 1-watt full scale to 500-watts full scale without changing plug-in elements. This universal flexibility is accomplished by eight easily switched ranges: Four forward power levels (10-500 watts) and four reflected power values (1-50 watts). The lower reflected power ranges are also available for forward readings by reversing rf connections.

In operation, a precision machined 50-ohm reference line-section is inserted between the signal source and the antenna, load or other component under power test. Directional power sensors incorporated in this line-section produce dc signals proportional to both incident and reflected rf main-line power, for readout on scales calibrated in watts as well as dB. All variable measurement parameters — frequency range, forward/reflected power and full scale values — are switched right on the front panel. The read-out unit and the line-section may be separated by as much as 3 feet for operational convenience.

The new multi-range model 4372 Thruline Wattmeter is ideal for CW, a-m, fm, ssb and tv signals, for design and maintenance of oscillators, transmitters and transceivers, for antenna matching, and design and development of rf components such as filters, sensors, loads and

More Details? CHECK—OFF Page 94

power transistors. It requires neither ac nor battery power.

Measurement accuracy is ±5% on all ranges from 0-1 watt to 0-500 watts at 25-500 MHz. Model 4372 fits standard 19-inch racks and takes 5¼-inches of panel width. For more information, write to Bird Electronic Corporation, 30303 Aurora Road, Cleveland (Solon), Ohio 44139, or use *check-off* on page 94.

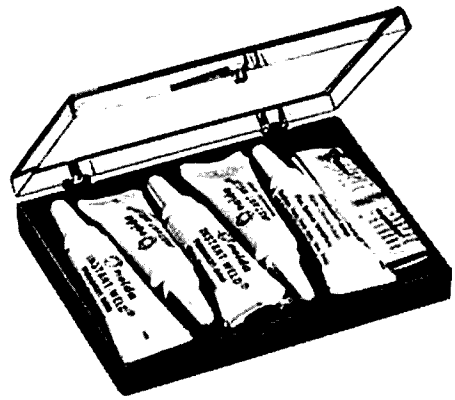
automatic six-channel scanner

Hamtronics has announced a new scanner kit to help vhf amateurs and monitor buffs complete the project presented on page 22 of the February, 1973, issue of *ham radio* magazine. The scanner is compatible with any crystal-controlled receiver with a squelch circuit as described in K2ZLG's article. Although designed primarily to operate with the Hamtronics Mark II FM Receiver and Six-channel Adapter, it is relatively easy to convert the oscillator circuit of any receiver or the receiver oscillator section of any transceiver to diode switching to allow the scanner to automatically crystal-switch. Channel indication may be provided, if desired, by a digital readout or discrete lamps. Full conversion instructions for various sets are given in the magazine article.

The kit comes complete with a 3-3/4 x 3-3/4-inch etched pc board (undrilled), all integrated circuits, transistors, resistors, isolation diodes, zener diodes and capacitors required to complete the unit. Also included are a schematic diagram and a pc board layout diagram. Price is $14.95 postpaid anywhere in the U.S. or Canada.

Available accessories include a seven-segment digital readout, priced at $5.00, and a number-66 twist drill (use in drill press to avoid breaking), $0.65. Send orders with remittance to Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612. Orders shipped promptly on receipt. For faster delivery, add 50 cents for air mail postage differential. For more information, use *check-off* on page 94.

instant-weld evaluation kit



Oneida Electronics has recently made available a new five-pack selection of its permanent bond *Instant-Weld* alpha cyancacrylate adhesive. The powerful, single component adhesive is said to have a tensile strength up to 5,000 pounds per square inch in each drop.

Four different formulas are available to provide users with the one best suited for the types of materials they want to bond. The five-pack Evaluation Kit contains all of the types, conveniently packed in 2-gram tubes totalling up to 760 one-drop applications. Red Label 101, for use when bonding any combination of plastic, rubber, ceramic or glass; sets in 10 to 20 seconds. Blue Label 102, for use when bonding any combination of plastic, rubber, ceramic or glass; sets in 45 to 60 seconds. Yellow Label 747, for use when metal is one or both of the bonded components, such as metal to plastic or metal to rubber, glass, etc; sets in 30 to 45 seconds.

Green Label 240, holds porous and nonporous materials. (Other grades of *Instant-Weld* bond non-porous materials — metals, plastics, glass, porcelain, hard woods, rubber, etc.) *Instant-Weld* 240 holds all of these materials plus leather, paper, soft woods and many fibrous substances; sets in 60 to 120 seconds.

The kit has been made up to help users find the most effective type for their specific bonding application. It's available from Oneida Mfg., Inc., Meadville, Pennsylvania 16335 for $7.95 postpaid. For more information, use *check-off* on page 94.

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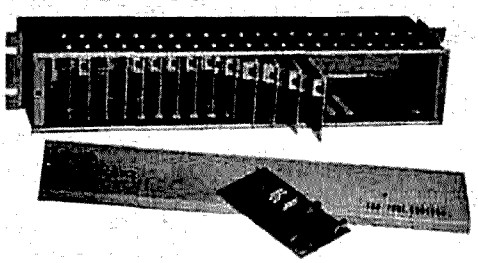
Dept. E 925 Sherwood Drive Lake Bluff, Ill. 60044

scanner-monitor servicing data

The new Howard W. Sams publication, *Scanner-Monitor Servicing Data,* provides up-to-date servicing data on vhf fm scanners and monitors. This new book features the most comprehensive gathering of service data for 30 of the most popular vhf and uhf receivers in the field, including such well-known trade names as B&K, Browning, Johnson, Midland, Pace, Pearce-Simpson, Penneys, Realistic, Sonar and Teaberry.

Scanner-Monitor Servicing Data provides valuable up-to-the-minute servicing information in the form of schematics, voltages, alignment, parts lists, crystal formula data, pictorial presentations, general troubleshooting information, etc. This handy new manual is priced at $4.95 postpaid, from Comtec Books, Greenville, New Hampshire 03048.

repeater tone panel



Alpha Electronics has announced its new multi-frequency repeater tone panel for use on repeater systems where tone control of numerous functions is required. Replacing the popular RCP-760 and RCP-770 tone panels, the new Alpha RCP-780 is considered the most modern, universal and compact tone control panel available. The unit is capable of handling up to 18 separate tone-controlled functions, using a modular plug-in card for each function.

A variety of tone system configurations are possible with the RCP-780. System A will respond to a received tone, either sub-audible or pulse tone, by keying and modulating the transmitter with the same tone, filtered and regenerated. System B removes the received tone from

More Details? CHECK–OFF Page 94

the audio and keys and modulates the transmitter with the received signal minus the tone. A new tone is generated for the transmit carrier which can be the same tone frequency as the received tone or may be a different frequency if desired. System C utilizes any number of pulse tone combinations to accomplish the same action as system B.

All tone encoding and decoding circuits use thick-film hybrid chip modules that plug into the individual carrier cards. Each tone card plugs into a gold-plated connector and all adjustments and indicator lights are at the front edge of the card for easy access.

Optional accessories include an adjustable time-out timer, carrier operated switch, wire line termination panel to allow hard wire interconnect, and low standby current drain option, expansion to accomodate 56 tone frequencies and provision for input-output relay drive circuits to control various functions.

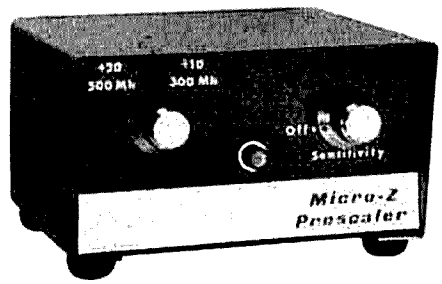
For additional information write to Alpha Electronic Services Inc., 8431 Monroe Avenue, Stanton, California 90680, or use *check-off* on page 94.

audio frequency meter



The new 1200A audio frequency meter from Linear Digital Systems incorporates the latest LSI integrated-circuit chips and a solid-state LED display that account for its small size and excellent reliability. It was designed to withstand the abuse found in university laboratories and in the field. A crystal-controlled

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The model 1200A audio frequency meter is priced at $245. For more information, write to Linear Digital Systems, Box 954, Glenwood Springs, Colorado 81601, or use *check-off* on page 94.

etco electronics catalog

In this day and age of electronics catalogs showing only lifeless mixes of fast-moving merchandise, the *ETCO Idea Book* comes as a refreshing surprise. This 1973 catalog is the outgrowth of ETCO founder Marvin Birnbom's (VE2ANN) desire to provide a little entertainment, some mailorder browsing, a little education, hundreds of surplus bargains, along with pages and pages of hard-to-find parts, gadgets and unusual items.

This fascinating 126-page catalog is bound to excite the imagination of everyone from the 9-year-old budding electronic genius to the 90-year-old, young-at-heart, engineer and inventor. The $1.00 (redeemable) requested to cover costs is well spent. Mail name and address and $1.00 to ETCO Electronics, Box 741, Montreal, Canada.

More Details? CHECK–OFF Page 94

bleeptone



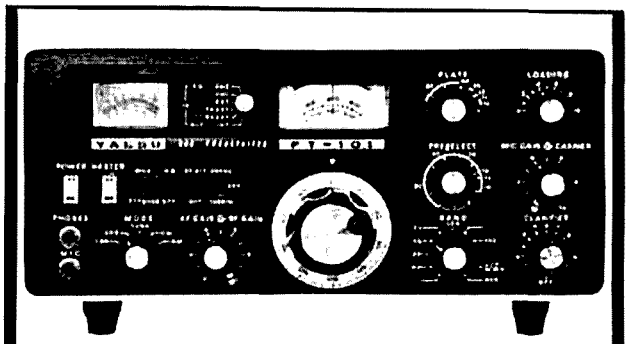
The Bleeptone compact audible signal source can be used as a code practice oscillator, circuit continuity tester, null detector or audible alerting device. The Bleeptone emits an audible signal of 70 dB to 86 dB sound pressure level at one meter when 8 to 16 Vdc is applied to its terminals. The very low current drain is 5 to 9 mA. No rfi is produced. The Bleeptone is available with standard nominal frequencies of either 2.5 kHz or 1.0 kHz.

Both a nylon ring and a horn adapter are available for mounting the Bleeptone. For additional information, please contact Cybersonic Division, C.A. Briggs Company, Box 151, Glenside, Pennsylvania 19038 or use *check-off* on page 94.

mobile antenna mount



A new mobile antenna mount that features fewer parts, simplified and quicker installation, lower silhouette and positive weather protection has just been introduced by Larsen Electronics. Designed to go into the usual 3-inch hole, the mount is adaptable to any location on



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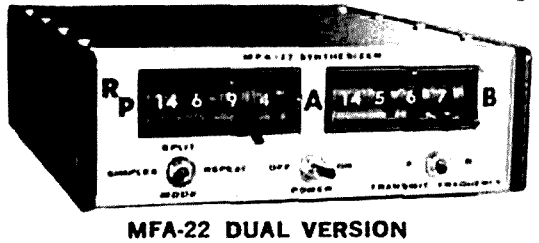
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To install the Larsen mount you only have to attach the coax according to the step-by-step-full-scale illustrated instructions which are included. The under part of the mount goes easily through the ¾ inch hole and a tough plastic fitting and weather-proofing O-ring spins into place. With those three steps the mount is ready to receive the antenna.

For more details write to Larsen Electronics, Inc., Box 1686, Vancouver, Washington 98663 or use *check-off* on page 94.

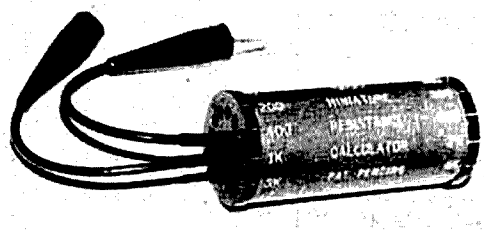
frequency counter

Miida Electronics announces the introduction of its new Digipet-60 Frequency Counter. This frequency counter has a *range of 1 kHz to* 60 MHz, extendable to 160 MHz with the Digipet 160 converter. It features a resolution of 1 kHz to 1 Hz (at 1 ms or 1 second gate times). It can be operated on either ac or dc, with complete overload protection, plus a stability aging rate of one part in a million per week. The entire unit is a compact 7-inches deep by 2½ inches high. It sells for $299.

For further information contact Andy Babiak at Miida Electronics, Division of Marubeni America Corporation, 2 Hammarskjold Plaza, New York, New York 10017 or use *check-off* on page 94.

More Details? CHECK–OFF Page 94

multifunction resistor decade



Called the R-1 Miniature Resistance Calculator by the manufacturer, Lee Electronic Labs, this handy new unit serves as a 10% resistance decade, a voltage divider or a resistor substitution box. Values between 100 ohms and 11 megohms are possible. Short leads plug into different jacks on the unit allowing 825 different values without the need for sequential switches and the inevitable wear which comes with their use.

The new unit is small, self-contained and lightweight. It comes with short leads to facilitate use in rack-mounted gear and in equipment where long leads to a decade box might introduce hum and undesirable rf pickup.

The unit sells for $17.45, postpaid. Complete specifications are available from Lee Electronic Laboratories, Inc., 88 Evans Street, Watertown, Massachusetts 02172 or by using *check-off* on page 94.

confidential frequency list

A *Who's Who* of unusual radio stations has just been published by Gilfer Associates. Called the "Confidential Frequency List," this book is the first major compilation of a-m, CW, ssb, RTTY and FAX nonbroadcast stations made available to the general public. Frequencies, callsigns, locations, schedules and radiated power are shown for thousands of radio stations operating between the broadcasting and ham bands from 12 to 27,240 kHz.

Prepared by Robert B. Grove, WA4PYQ, the "Confidential Frequency List" reveals radio frequency and callsign

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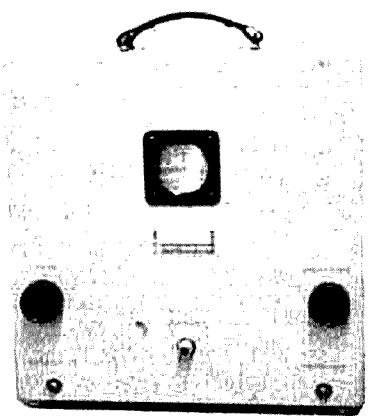
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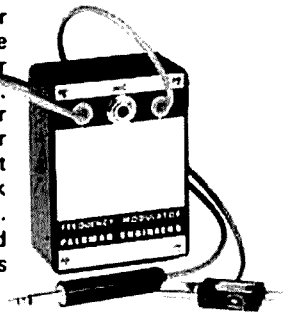
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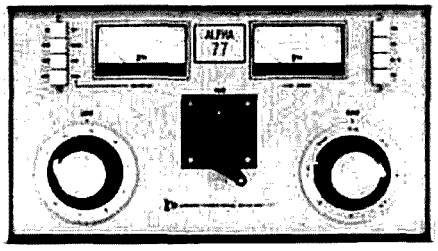


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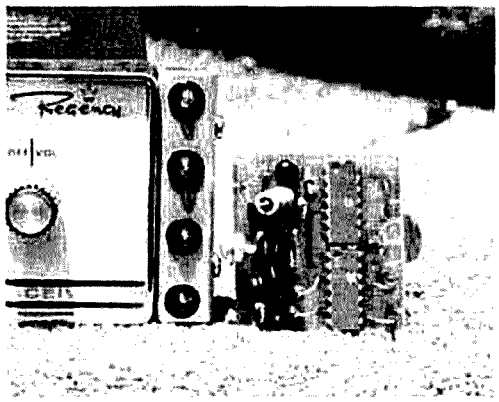
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information heretofore kept under wraps. There are 34 separate listings in the list including: Interpol, CIA, RTTY Press, USAF Global Aero, spy and number stations, radio-beacons, weather broadcasters, Flying Doctor Service, foreign embassy networks and hurricane hunters.

The "Confidential Frequency List" is one of a series of new publications from GILFER Associates, specialists in disseminating information on radio frequency usage around the world. 64 pages. Soft-bound. $3.95 from Gilfer Associates, Inc., Box 239, Park Ridge, New Jersey 07656.

regency scanner modification



The popular Regency HR2, HR2A and HR212 two-meter fm transceivers can be made to scan four channels with Diamond Enterprise's model 4CS scanner. The 4CS comes as a kit for $29.95 or ready-built for $49.95. Both units use a fiberglass circuit board and come with all mounting hardware and an external light-emitting diode display. The display is small – only ½-inch wide and just as high as the Regency transceivers. The scanner board itself is also small – 1 5/8 by 1½ inches – and fits inside the transceiver cabinet.

The units allow the transceiver to scan four different channels – with the LED display indicating which channel is being monitored. The unit is activated by the transceiver channel position switch.

For more information contact Triangle Enterprises, 404 North Centennial Street, Indianapolis, Indiana 46222 or use *check-off* on page 94.

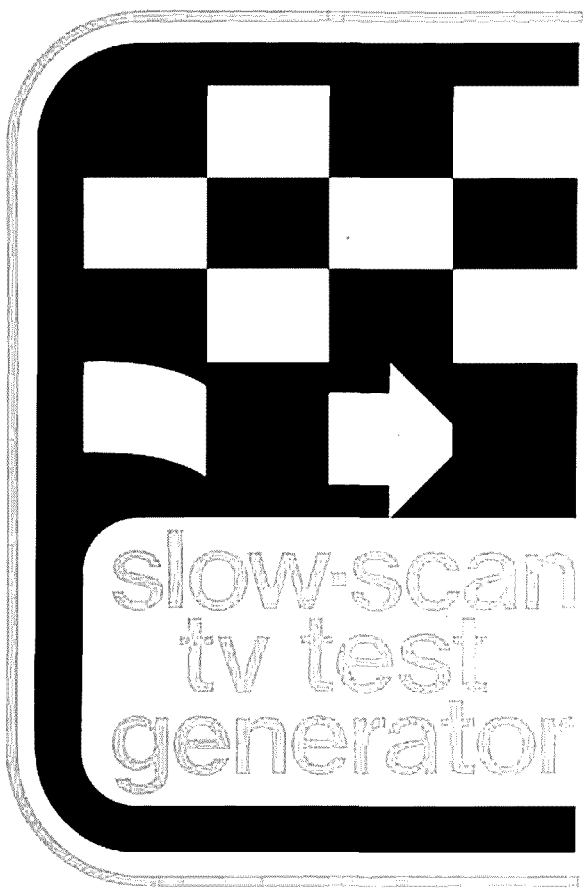
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ham radio magazine

JULY 1973



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July, 1973
volume 6, number 7

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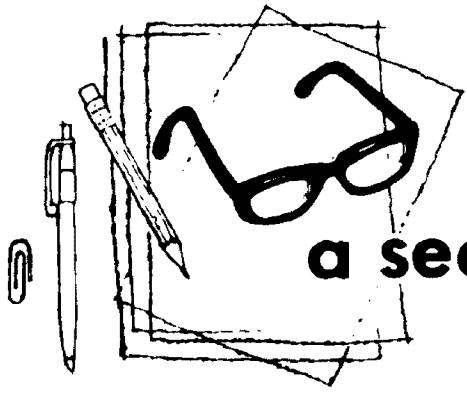
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a second look

by jim fisk

Amateur radio operators, especially those on two-meter fm, are using more and more dry batteries than ever before. Zinc-carbon batteries rate very high on the list because they are relatively inexpensive and easy to find, although some amateurs use the more expensive but higher powered alkaline-manganese cell, and a few swear by rechargeable nicads. A new dry battery, which will be on the market in the near future, just might revolutionize the whole field of portable dc power.

The new battery, the lithium organic cell, has been receiving enthusiastic reviews from the military, which depends on man-carried batteries for much of its power. The reason for the enthusiasm is that lithium apparently produces a battery with greater energy density than that of any other existing type. Lithium batteries are lighter, have greater power output, can operate over wide temperature ranges and have a remarkably long shelf life -- up to 20 years.

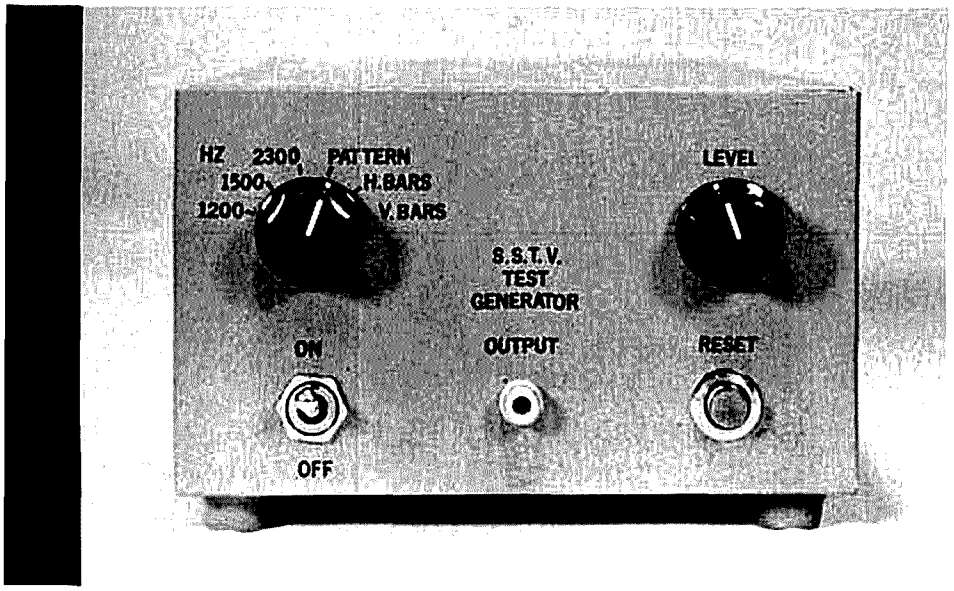
Several companies are currently manufacturing lithium cells, including Mallory, Eagle-Picher and Power Conversion, Inc. Both Mallory and Power Conversion use lithium with a sulfur dioxide electrolyte; experimental Eagle-Picher lithium batteries are based on lithium and a carbon flourine compound in conjunction wtih an organic electrolyte.

Mallory also has a solid-state lithium battery that uses a metal lithium anode and a metal salt as the cathode. The electrolyte is an electronically insulating solid. Because of the reactive nature of these materials when exposed to the atmosphere, this battery must be hermetically sealed. However, the absence of any liquid in the battery completely eliminates any corrosion or gassing. In fact, these cells have been stored for long periods at more than 200° F with no detectable loss in energy capacity.

One of the big advantages of the new lithium batteries is their very high energy density. Prior to this, the most energetic batteries have been the silver-zinc units used on the manned Apollo program -- they provided approximately 110 watt-hours per pound per cell. Some of the new lithium batteries can generate 200 watt-hours per pound per cell, a nearly 85% increase. When compared to carbon/zinc and other commonly used batteries, the energy density of the lithium cell is even more impressive. What this means, basically, is that if you presently use 5 pounds of batteries to power your communications gear, lithium batteries would cut the weight in half, approximately.

As far as power output is concerned, the energy from one lithium D-cell at a discharge rate of 1 ampere is equivalent to four mercury-zinc cells, five alkaline-manganese or 30 carbon-zinc cells! The introductory cost for these new lithium batteries is expected to be quite high, about $9.00 each for D-cells, but that price can be expected to come down as usage and production increase. However, when you consider that one lithium D-cell can provide the same power as 30 carbon-zinc units, the price isn't nearly as astronomical as it first appears. Now, if somebody can figure out a way to recharge them. . .

Jim Fisk, W1DTY
editor



slow-scan tv test generator

Complete construction details for a slow-scan tv pattern generator for troubleshooting sstv equipment

Bert Kelley, K4EEU, 2307 South Clark Avenue, Tampa, Florida 33609

Anyone who builds his own slow-scan television equipment soon learns why so many other operators go the commercial equipment route. It's not so much the circuit complexity or parts procurement problems as it is the difficulty of getting the circuits properly adjusted. Almost all the circuits in the sstv monitor have to be operating correctly before you can see any results on the picture tube, so the monitor itself is useless as a test unit.

Some authors have suggested that a tape recorder and a length of audio tape be used as a test signal during adjustments. However, unless this tape is obtained from a sstv manufacturer, the quality is questionable. The sync and video pulses may be off frequency because it is almost impossible to accurately tune in a sstv signal without a monitor. To add to the difficulty, video quality will vary from station to station, and there is usually interference from nearby stations. Any good transmissions will be short so the tape will have to be rewound again and again during use.

What is needed is a sstv test generator which can deliver a continuous high quality test signal and require no atten-

tion. The unit described here can be used to align and test bandpass filters, video discriminators, video amplifiers and sync separators, as well as to check horizontal and vertical sweep linearity and size.

features

If test equipment is to be useful, it should be accurate. The sync and video frequencies in this sstv test generator are inherently accurate because they are derived from sources that do not vary. Digital logic does the rest. The division ratios are set by selection of IC types and the pin-to-pin strapping. There are no multivibrators that have to be locked in with a pot adjustment.

nominal 1200 Hz. This means that the sstv circuits may be adjusted with confidence in your signal source.

test signals

A checkerboard test pattern was chosen as the primary test signal because the deep transitions between peak white and picture black will critically test any sync separator circuits. The 60-Hz square wave "video" is an easily recognized pattern that can be traced through the video circuits of the sstv monitor with any scope.

Since the transitions are sharp and clean, the reproduced picture should have sharp, clean edges and this gives an

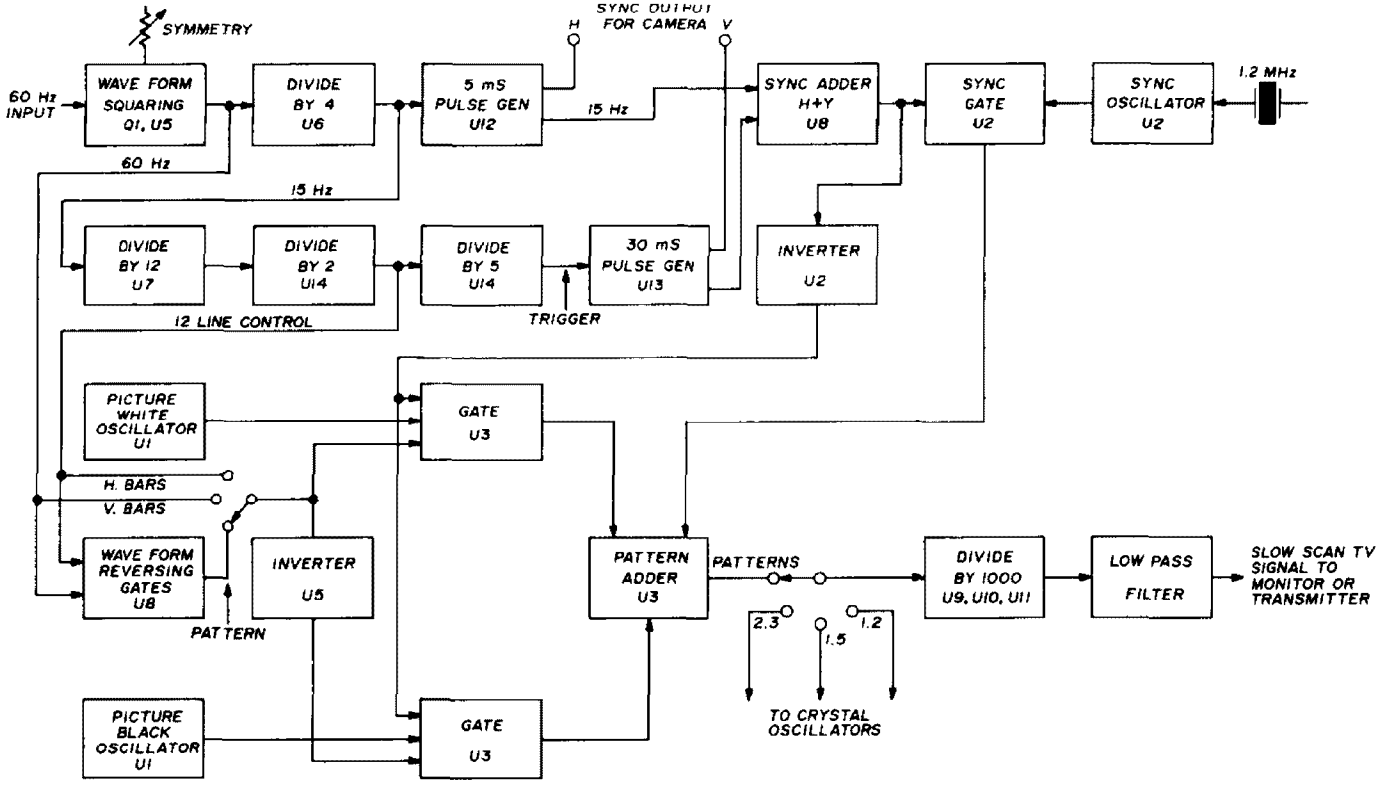


fig. 1. Block diagram of the slow-scan tv test-pattern generator.

The three basic frequencies are derived from crystal oscillators and divided down to the sstv range with digital logic. Digital division removes switching transients and increases the accuracy of the audio tones. The 1.2-MHz crystal may be as much as 1 kHz off frequency, but the sstv sync frequency will still be within 1 Hz of the

excellent indication of picture resolution (see checkerboard photo).

The horizontal and vertical bars are a good indication of sweep linearity although the checkerboard pattern could also be used for this purpose. Since the horizontal and vertical signals are already available in the circuitry, it was only a

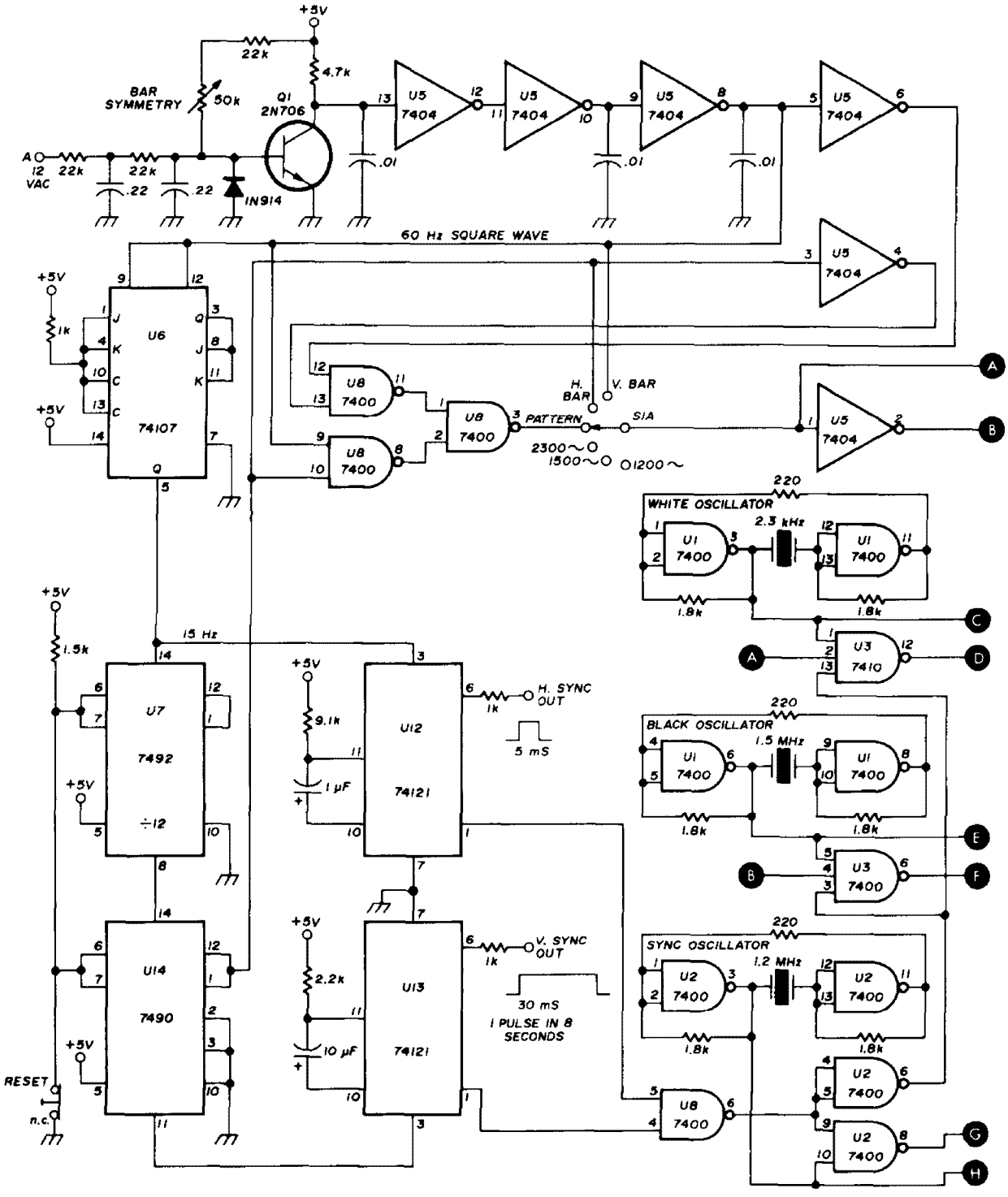


fig. 2. Circuit for the sstv test generator makes extensive use of integrated circuits. Complete unit is housed in a small 3-1/2x6x10-inch Minibox.

matter of adding a selector switch to provide three video patterns as well as the three basic slow-scan audio tones of 1200, 1500 and 2300 Hz. These tones can be used to tune discriminator coils or the all important sync-separator circuit.

how it works

Refer to the block diagram in fig. 1 and the logic diagram in fig. 2. The 60-Hz signal that drives the test generator is obtained from one side of a 24-volt center-tapped power transformer, filtered, and connected to a transistor, Q1, which turns on for each positive going half-cycle of the waveform. The bias on Q1 is adjusted by a 50k trimmer to make the output symmetrical. The signal is

further squared by three sections of a SN7404 integrated circuit, U5.

The square-wave output from U5 is applied to a JK flip-flop, U6, which divides by four. This is the slow-scan horizontal scanning frequency. However, the pulses are too long so they are used to trigger a 5-millisecond pulse generator, U12. Each pulse generator IC has two outputs. One output is normally low and one is normally high. The positive-going pulse at the Q output is connected to the rear panel jacks for external use.

Two cascaded IC counters, U7 and U14, count the 120 horizontal lines and trigger the 30-millisecond vertical monostable oscillator, U13. Horizontal and vertical sync pulses are then combined in the sync adder, U8. The sync signal is connected to signal gates so that sync always has precedence.

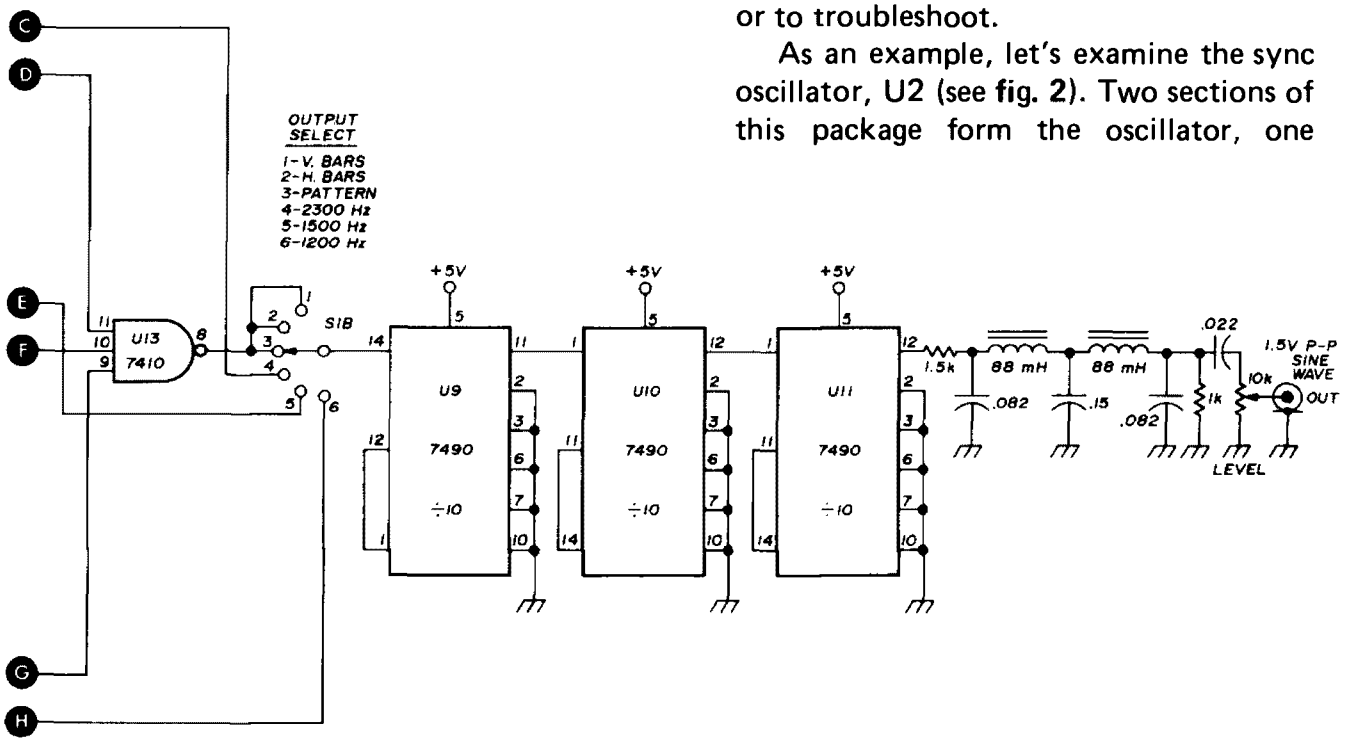
The three key frequencies in the generator are provided by continuously operating crystal oscillators that generate rf square waves. These three oscillators are gated so that only one signal may be on at a time. The output is a sequentially switched rf waveform divided down to the audio range by U9, U10 and U11.

The lowpass audio filter* does a good job cleaning up the square wave output from U11. The measured harmonic distortion is only 2.4% at 1200 Hz, 0.53% at 1500 Hz and 0.1% at 2300 Hz.

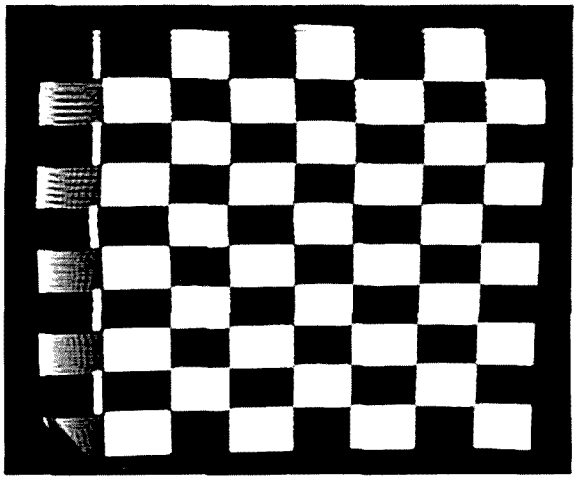
the digital logic

The TTL logic used in this circuit is fairly fast — much faster than needed. This can be a problem if the filtering is omitted at the base of Q1 or the input to U5 — the unit would become sensitive to power line spikes and result in erratic operation. This logic, however, provides at least two advantages. First, the signal frequencies used in this generator are mostly dc or 60 Hz so any small oscilloscope may be used. Also, most of the circuits are ordinary simple TTL NAND gates, so they are not hard to understand or to troubleshoot.

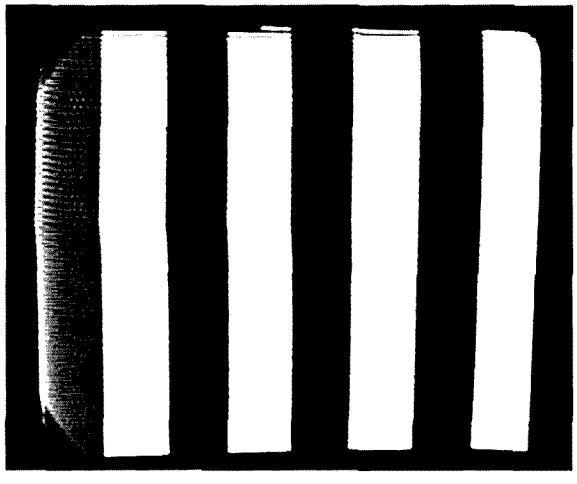
As an example, let's examine the sync oscillator, U2 (see **fig. 2**). Two sections of this package form the oscillator, one section is used as a gate, and one as an inverter. A NAND gate has to have all inputs positive to get a low, or zero, volt output. If pins 4 and 5 of U2 are connected together the gate is converted to an inverter. If, on one section of the



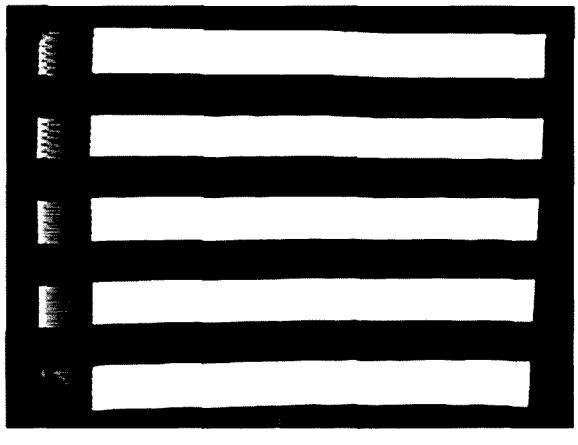
*The 88-mH toroids used in the low-pass filter may be obtained from M. Weinschenker, Box 353, Irwin, Pennsylvania 15642, five for $2.00, postpaid.



Checkerboard test pattern displayed on the K4EEU sstv monitor. The gray bars on left are due to second harmonics of the sync frequency, removed by the test-generator filter, but restored by the limiter in the monitor. This second harmonic, at 2400 Hz, is close to 2300 Hz, sstv picture white.



Vertical-bar test pattern displayed on the sstv monitor. Gray bar on the left side of the monitor tube is caused by the second harmonic of the sync frequency.



Sstv horizontal-bar test pattern, displayed on K4EEU's monitor.

IC, the output is connected back to the input with a resistor, an inverse feedback connection is made which linearizes the gate to a degree and prevents crystal lockup or starting problems.

The output at pin 3, U2, is a 3.5-volt p-p rf square wave, or, more accurately, it is a pulsating dc voltage which rises in a positive direction from the zero baseline up to 3.5 volts and falls again to zero at an rf rate. If this square wave is applied to pin 10, U2, it will pass through, inverted, to pin 8 provided the gate is enabled by making pin 9 positive. Normally, however, pin 9 of U2 is at zero volts and the gate is inhibited. Therefore, the output at pin 8 is high and stays high regardless of what signal is applied to the input, pin 10. To turn off any input to a NAND gate, any of the other inputs can be made low.

The sync-pulse monostable outputs at pin 1 of U12 and U13 are normally high; both are connected to the sync adder gate, U8. With both inputs high, the output at pin 6 of U8 is low. If either input to this gate goes low, as during a horizontal or vertical sync pulse, pin 6 goes high for the duration of that pulse. This enables pin 9, U2, and the sync rf passes through the gate.

When any gate is cut off, the output is high, so several such gates may be connected to a mixer such as U3 without inversion. When U2, pin 9, is high, turning on the sync, U2, pin 6, is low, or inverted, which inhibits the other two gates and prevents picture frequencies from passing during the sync interval. Pin 2, U5, is connected to pin 4, U3, so that either the black or white picture gates may be enabled, but not both at the same time.

The *waveform* reversing gate, U8, reverses the phase of the 60-Hz square wave every twelve picture lines. The output appears alternately at pin 8 or pin 11 of U8, but not both at the same time.

construction

Most of the components are mounted on a 5½ x 8-inch circuit board with the

rest of the parts mounted on the ends of a small 3½ x 6 x 10-inch Minibox. A LM309K voltage-regulator IC is directly bolted to the rear of the Minibox, which is used as a heat sink.

Complete information on board layout is given in **fig. 6**. The circuit trails are drawn on the copper with a RMP-700 Kepro resist pen, or one similar. A parts list and epoxy, plated circuit board is available from the author for $10.00 postpaid in the United States.*

Inexpensive crystals are available from JAN crystals from stock in the FT-243 holder, but I prefer the HC-6/U crystal holder since it takes less space on a printed-circuit board. The tolerance of the crystal is not critical; however, 0.005% tolerance custom ground crystals can be obtained.

Since this is a fairly complex construction project, I always build in sections and debug if necessary, starting at the power supply, **fig. 3**. Make sure the input of the LM309K voltage regulator is properly connected before turning on the supply. Verify the 5-volt output before wiring in any of the TTL devices (it is possible to have 18 volts unregulated where there should be only 5 volts). The regulator needs no adjustments and has built-in current limiting so that, if the 5-volt line is shorted, the current is held to about 1 ampere, protecting the power supply.

The sync generator section can now be built and checked out with an oscilloscope. There should be a square wave at the collector of the transistor Q1 at the input to U6, pins 9 and 12. The pulse width at pin 6, U12, should be 5 milliseconds. This can be measured with suf-

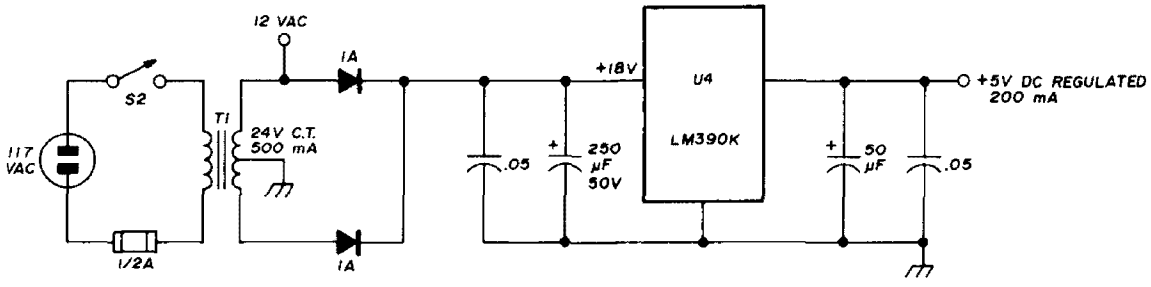


fig. 3. Power supply for the sstv generator. LM309K voltage-regulator IC uses one end of the Minibox enclosure for a heat sink.

*A kit of parts is available from Truman Boerkoel, K8JUG, Stotts-Friedman Company, 108 N. Jefferson Street, Dayton, Ohio 45402.

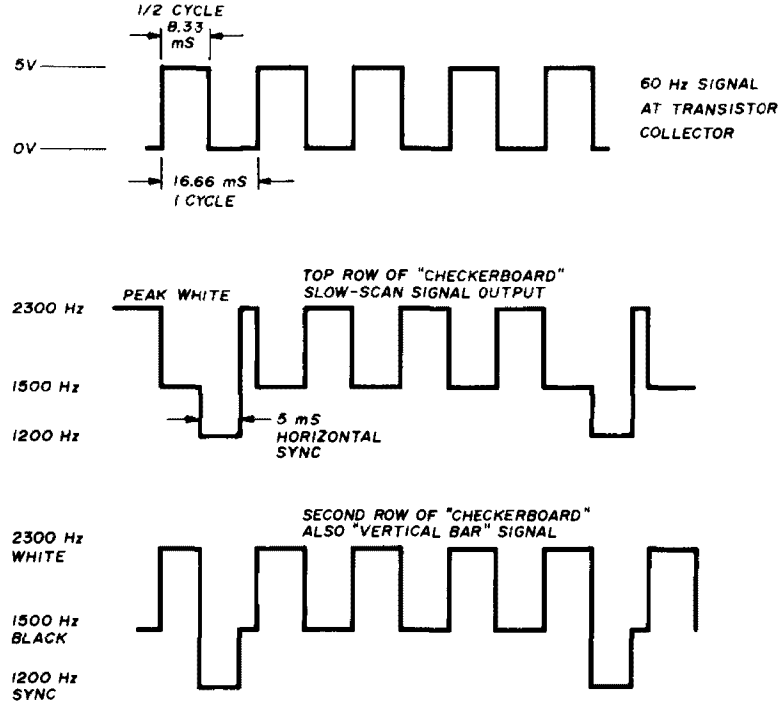


fig. 4. Output waveforms of the slow-scan television test generator.

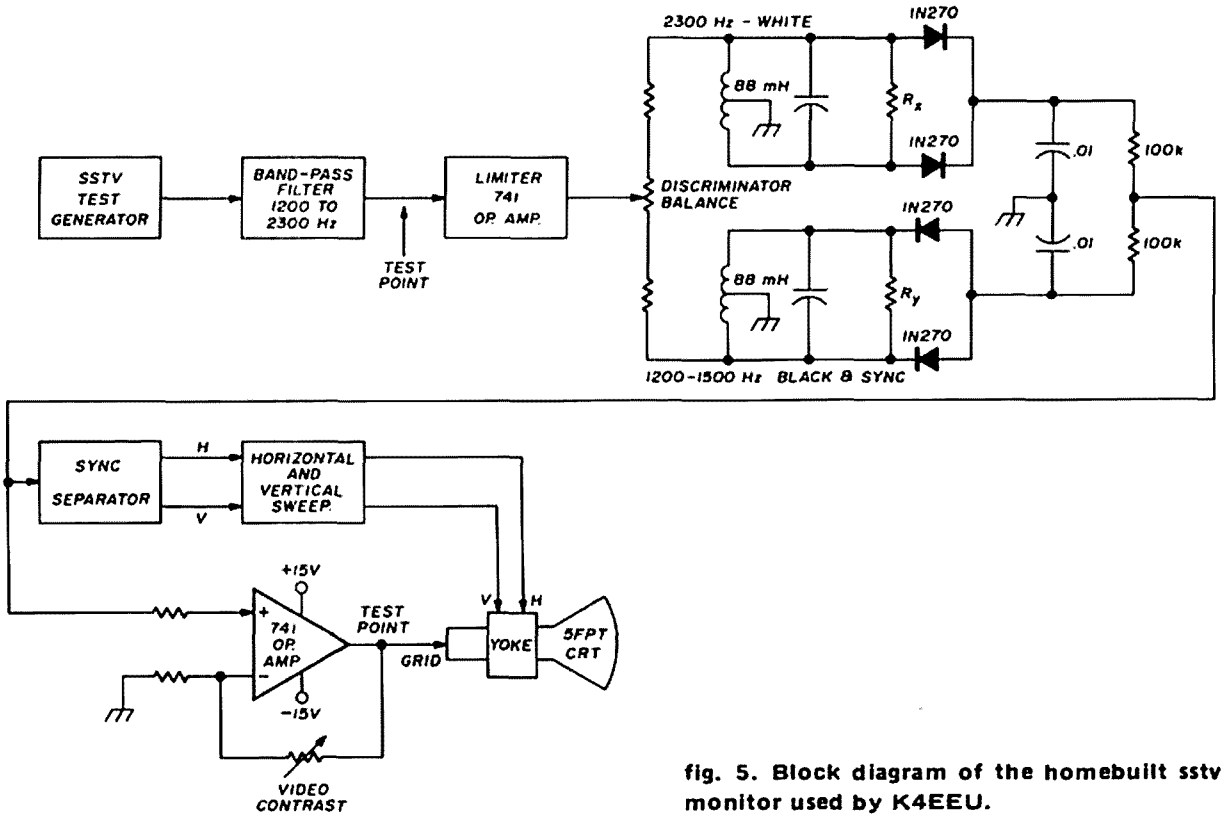
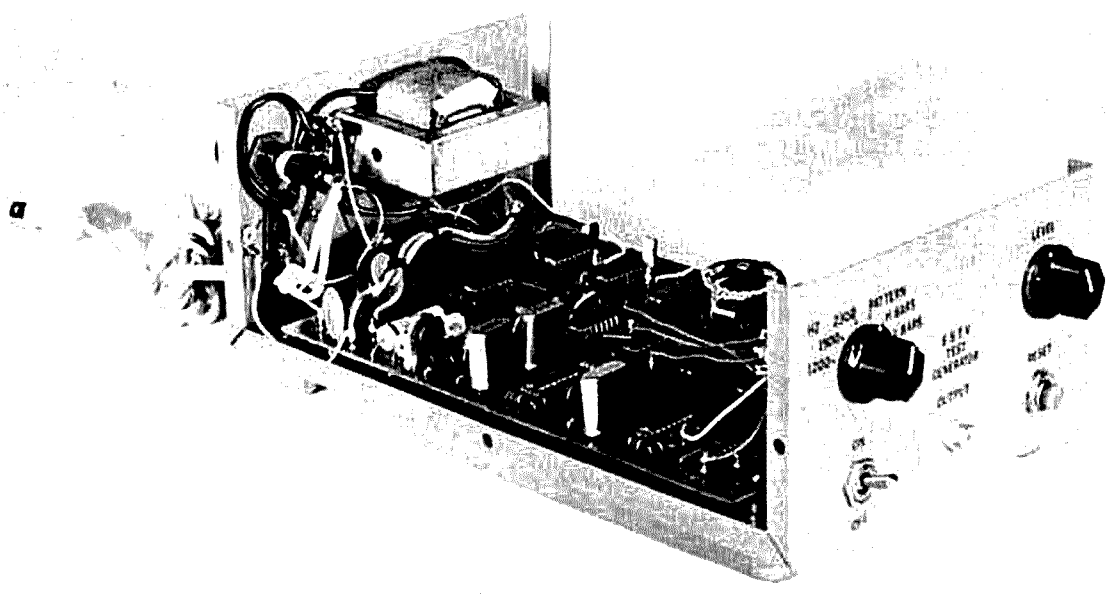


fig. 5. Block diagram of the homebuilt sstv monitor used by K4EEU.

ficient accuracy by referencing the scope to one cycle of the 60-Hz signal which has a time interval of 16.66 milliseconds. The sync pulse width should be about one-third the length of one 60-Hz cycle (see fig. 4).

The vertical sync pulse is a little more difficult to measure because it occurs once every eight seconds. This may be speeded up, either by repeatedly operating the *reset* button or by *temporarily* rerouting the lead that runs from pin 3, U13, to U14 so that it is connected directly to pin 5, U6. The pulse from U13 will then be at a 15-Hz rate and much easier to measure on a scope.



Construction of the slow-scan tv test generator.

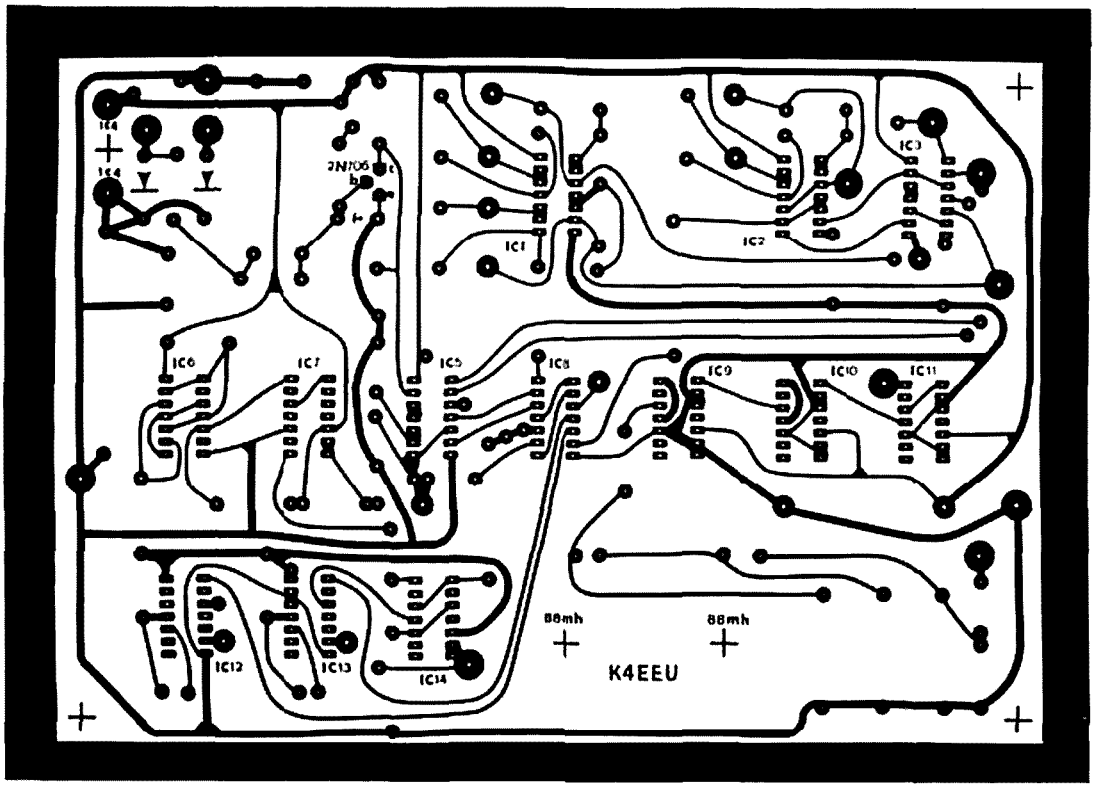


fig. 6. Layout for the printed-circuit board for the sstv test generator. A circuit board and parts list is available from the author.

The pulse width should not be less than 30 milliseconds. Two 60-Hz waves are about 33-milliseconds long and this is close enough for practical purposes. The pulse widths should be sufficiently close without any circuit changes. However, if they are not, the pulse width can be adjusted by changing the value of the resistor connected to pin 11 of the appropriate 74121 IC.

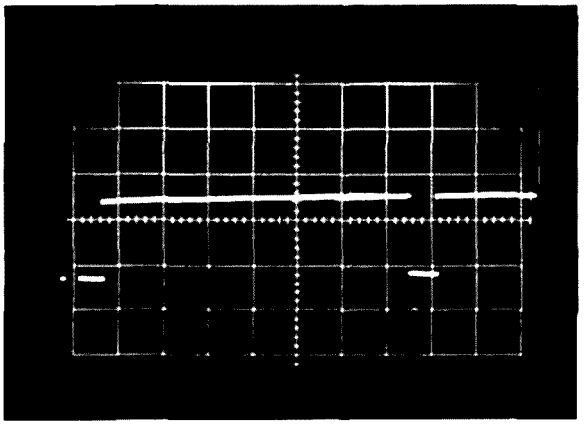
The pattern generator can be checked out by wiring in the waveform reversing gate, U8, and checking the waveform at pin 3. Externally sync your scope and see if the polarity of the waveform is reversing about once a second.

If you cannot observe the crystal oscillators on a scope, the signal can be detected on a broadcast receiver. Of course, if you have an output at the filter, it's a good indication that the crystal oscillators are operating all right.

The foregoing is not intended to emphasize the complications but to be of assistance if troubleshooting becomes necessary. Usually ICs can be wired into a circuit as so many building blocks, and, if wired correctly, everything works.

operation

The sstv test generator can be used with an oscilloscope to check the slow-scan signal at the different points in the monitor. If you want to see if the input bandpass filter is operating correctly, put the test generator switch on pattern or bars and look at the output of the filter with a scope. The amplitude should be the same for the three frequencies. Similarly, the demodulated composite video signal can be seen at the discriminator



Clipped and restored horizontal sync pulse taken from the sync separator of the K4EEU sstv monitor.

output and traced up to the picture tube grid.

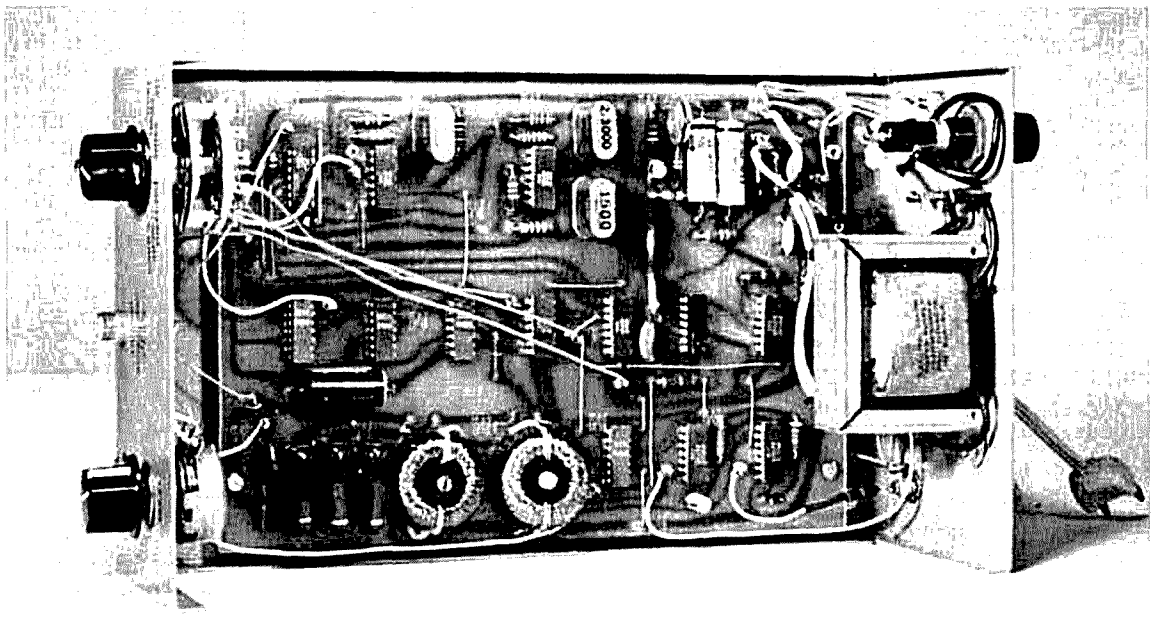
The sync separator circuit in the monitor should show a clean sync pulse with no interference between the sync and the black video pulses. This circuit can be optimized by setting the selector to 1200 Hz and tuning.

The sstv monitor may use a discriminator circuit similar to that used in a RTTY demodulator (fig. 5). The discriminator coils can be individually set on frequency with the test generator and tell about it by using variable pot (about 50k). The sync pulse waveform is relatively unimportant here but it should be separated from picture black. The sync circuit will clean up the sync pulse.

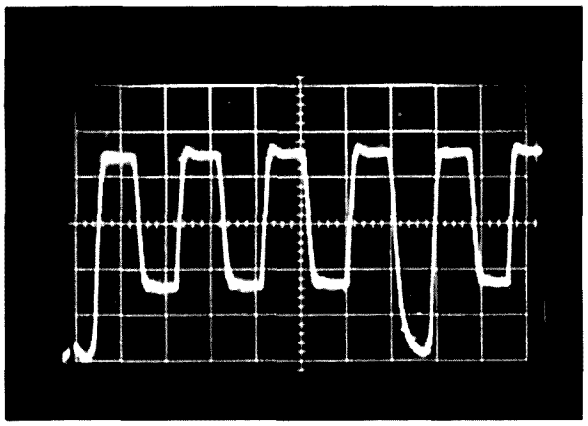
The reset button was installed on the test generator so that a vertical pulse could be obtained immediately at any time when working with a sync circuit in a monitor. Otherwise, the pulses are separated with an 8-second time interval. If this button is omitted, pins 6 and 7 of U7 and U14 must be grounded or the dividers will not operate.

Once you build this sstv test set, you will wonder how you ever got along without it. At my station it revived a faltering monitor project. With only a few hours work a trouble in the monitor sync separator was cleared and I started receiving good pictures.

then the exact value of the damping resistors selected by viewing the demodulated pattern on a scope. The correct resistor will result in minimum ringing and best square wave response, and can be found in less time than it takes to



Complete sstv test generator is built on 5-1/2x8-inch printed-circuit board and installed in a 3-1/2x6x10-inch Minibox. LM309K voltage-regulator IC is mounted on the rear of the Minibox.



Demodulated test-generator signal at the output of the monitor discriminator.

ham radio

operational-amplifier relay for motorola receivers

Complete construction details for the SOOPAR — a Signal-Operated Operational Relay

J. Jay O'Brien, W6GDO, 6606 Fifth Street, Rio Linda, California 95673

The SOOPAR, A Signal-Operated Operational Amplifier Relay, is the answer to your carrier-operated relay troubles. This circuit uses an inexpensive integrated circuit, an inexpensive, easy-to-obtain relay and a few parts to provide you with a reliable unit that can be adjusted to fit your needs for receiver control of an external device. The external device could be a light to indicate the receiver is receiving a signal, a Transmitter in a repeater system, or control functions such as a garage door opener, etc.

squelch circuit

The Motorola compensated squelch circuit, diagrammed in **fig. 1**, operates to silence the speaker when no signal is present. The SOOPAR derives its input signal from this squelch circuit.

The squelch circuit operation is as follows: Signal and noise from the antenna, mixers and the i-f amplifier stages are fed into the limiter stages which are driven into saturation, thus limiting the limiter output to a constant voltage. This output voltage, at the last i-f, is the same whether it is signal or noise or any proportion of both.

The action of driving the last limiter stage into saturation generates limiter grid current and, thus, a negative voltage at the limiter grid. This grid voltage is picked off, filtered and used as bias by the later squelch stages. The i-f output of the limiter is connected to the discriminator stage where it is converted to audio. The voice-audio components are amplified by the audio amplifiers and drive the speaker when the first audio stage is unsquelched.

Audio components produced by the discriminator that are above the voice-range (noise) are filtered out from the voice-range components and amplified by the noise amplifier. The noise amplifier gain may be adjusted by a variable resistor in the cathode circuit; this variable resistor is the squelch control.

The output of the noise amplifier is rectified by the noise rectifier. The positive-going dc output of the noise rectifier

is proportional to the noise output of the noise amplifier. That is, more noise into the rectifier produces a more positive dc voltage out of the rectifier.

The dc output of the noise rectifier is connected in series with the negative bias obtained from the limiter stage and is applied to the dc amplifier grid. The dc amplifier turns the first audio amplifier stage on and off. Typically, the dc amplifier keeps the first audio amplifier turned off (squelched) whenever its dc input from the noise rectifier is zero volts or more positive.

The dc amplifier turns the audio amplifier on when the dc voltage applied to it from the noise rectifier becomes more negative than zero volts. This condition occurs when the noise voltage is removed from the noise amplifier and rectifier. This "quieting" occurs when a carrier is received by the receiver.

This circuit has a measure of compensation built into the biasing arrangement. When the drive to the limiter stage is reduced as a result of tube aging or low supply voltage, the negative bias derived from the limiter is reduced proportionately. The noise component of the discriminator output is also reduced with tube aging or reduced supply voltage, resulting in a lower positive going noise rectifier output. Since these two voltages are of opposite polarity and are connected in series, the result is a negligible change in

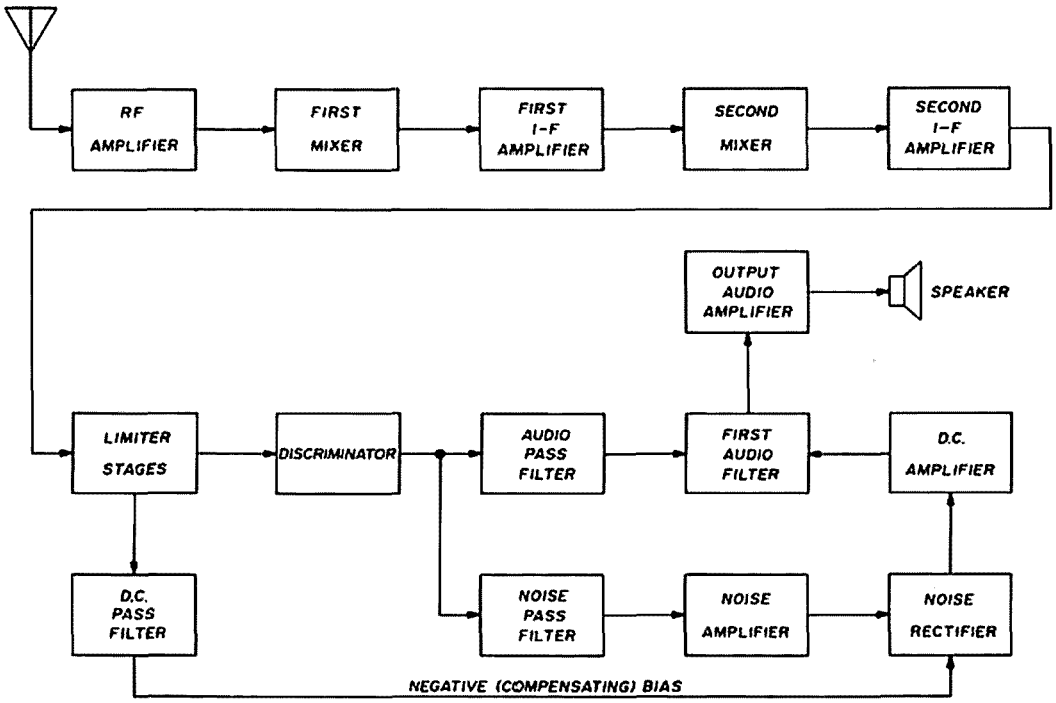
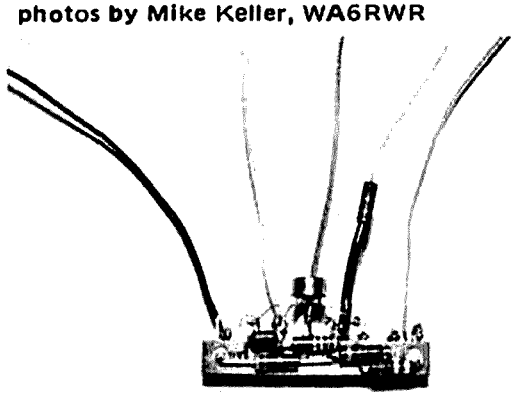


fig. 1. Block diagram of a typical Motorola receiver using a Motorola compensated squelch circuit.

photos by Mike Keller, WA6RWR



The SOOPAR, ready to be installed and connected to an external relay.

Editor's note: This is more than an article on a signal-operated op-amp relay — it also presents a comprehensive review of squelch operation in Motorola tube-type fm receivers.

the voltage fed to the dc amplifier from the noise rectifier. Thus, supply voltage changes and tube aging tend to be automatically compensated for by this circuit.

When the squelch control is set for reception of a slightly noisy but useable signal, the voltage presented to the dc amplifier when no signal is being received is approximately +5 volts.

tube-type carrier-operated relays

Tube-type carrier-operated relay (COR) circuits, when used, are connected to sample the voltage presented to the dc amplifier. A representative circuit is shown in fig. 2.

The signal voltage applied to the grid of the dc amplifier is sampled through R1 and connected to the grid of the first half of the COR tube which goes into conduction whenever the sampled voltage is more positive than the 1.8-volt cathode bias developed by R2 and R3. When the first half of the tube conducts, the voltage at its plate approaches zero. This is connected to the grid of the second half of the tube, which, under this condition, draws little current. The voltage drop across relay K1 is 27 volts or less, and K1 is released.

When the receiver is unsquelched, the voltage presented to the dc amplifier and sampled by the COR goes negative. This causes the current drawn thru R4 by the first half of the COR tube to reduce. As this current reduces, the voltage drop

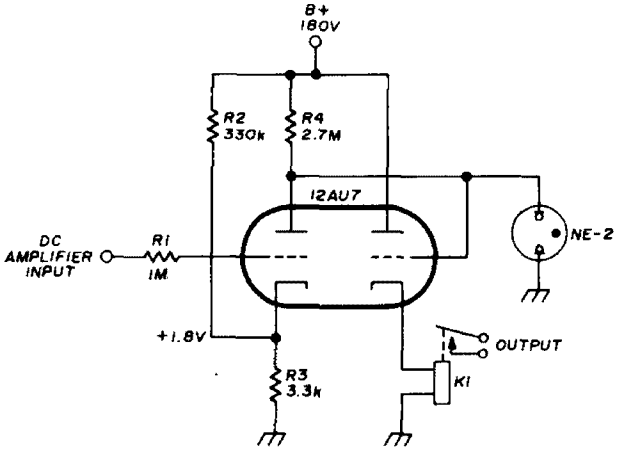


fig. 2. Motorola carrier-operated relay circuit. Relay K1 is an 8000-ohm relay which picks up at 39 volts and drops out at 27 volts.

across R4 reduces, making the voltage on the grid of the second half of the COR more positive. When this voltage rises to about 50 volts, the neon tube fires, preventing the voltage from reaching a higher potential that could damage the tube.

When the voltage at the second grid goes more positive, the second half of the COR draws more current through K1. When the current is sufficient to produce a 35-volt drop across K1, it closes, actuating the external device.

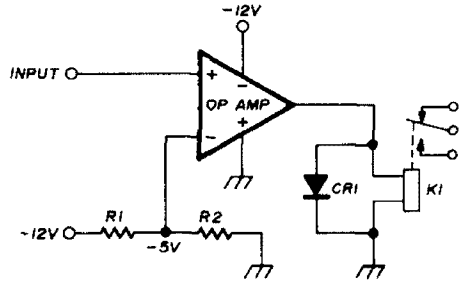


fig. 3. Circuit for using an operational amplifier IC to operate a relay.

cor disadvantages

Users of conventional tube type CORs will agree that the COR is the real trouble spot in a repeater receiver. Part of this is due to the fact that the voltage across the relay does not go from zero to "all on" when a signal is received. Rather, the relay operate and release voltages play an important part in COR operation. As the COR tube ages, the voltage presented to the relay changes, thus effectively changing the COR sensitivity. A circuit that would only produce two relay voltage conditions, full on and full off, would eliminate most of these relay problems.

Another problem is the inconsistent loading effect that the COR has on the receiver squelch circuit. When the squelch voltage is more positive than the voltage on the cathode of the first half of the COR tube, the grid and cathode act like a diode and conduct current, placing a 1-megohm load on the squelch circuit and reducing its adjustment range. Below this voltage, the current is negligible unless the COR tube becomes gassy; when this happens, the effect is unpredictable.

It is difficult to adjust this type of tube COR to open and close in the correct relationship to the audio squelch. The proper relationship is one where the audio squelch opens just before the COR closes as a received signal is slowly increased in level, and where the COR opens just before or at the same time as the audio squelch closes when the signal is decreased. This is almost impossible to maintain with a tube-type COR, even if an additional control is added to independently adjust COR sensitivity.

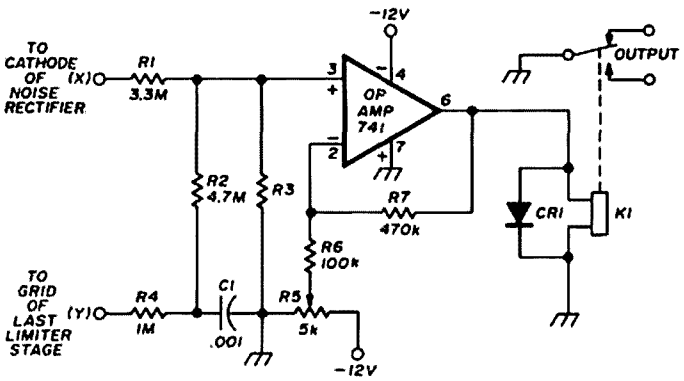


fig. 4. Schematic for the SOOPAR, signal-operated operational relay. Relay K1 is a Sigma 65FP1A. Diode CR1 is silicon. See table 1 for value of R3.

op-amps

Operational amplifiers are miracle devices that seem to perform magic. In the case of the SOOPAR, an op-amp is used as a dc amplifier with almost infinite gain. In this case the op-amp looks at two input voltages; one constant (adjustable) and one variable with signal level. When the variable input is equal or more positive in voltage than the constant, the output from the op amp is in the "off" condition.

When the variable input is slightly negative as compared to the constant, the output of the op amp changes to the "on" condition. A 10-volt change (from 2 to 12 volts) at the output is produced by an input voltage change of a hundredth of a volt or so. The op amp in the SOOPAR is capable of directly operating a relay; of course, this voltage change can be adapted to solid-state circuits that do not use any relays.

op-amp relay

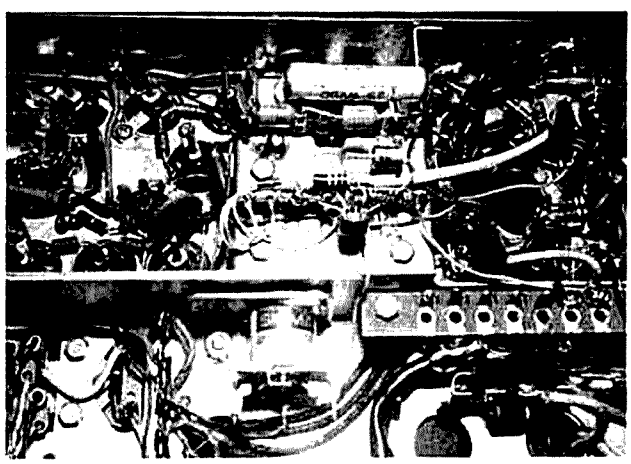
A simple circuit using an op-amp to sense an input voltage and operate a relay is shown in fig. 3. A negative 12-volt supply is connected to the op-amp and to a voltage divider, R1 and R2. The inverting input of the op-amp is connected to the junction of R1 and R2, whose values are chosen to present a bias potential of -5 volts. The non-inverting input of the op-amp is the input terminal of the op-amp relay.

If the input voltage is between zero and -5 volts, the relay is released. When the voltage is raised above -5 volts, the relay is operated. To prevent damage to the op-amp the input voltage should not exceed the supply voltages. The current drawn by relay K1 must not exceed the rating of the op-amp. Diode CR1 prevents damage to the op-amp from reverse voltage spikes generated by K1 when it is released.

soopar

The signal-operated op-amp relay or SOOPAR is an adaption of the op-amp relay circuit and is shown in fig. 4. Resistor R5 takes the place of R1 and R2 in fig. 3 and is set, nominally, to about -5 volts. R1 is connected to the squelch circuit of the receiver, and R4 picks off the limiter voltage. C1 and R4 act as a filter to remove the i-f signal voltage present at the grid of the last limiter.

Resistors R1, R2, R3 and R4 make up

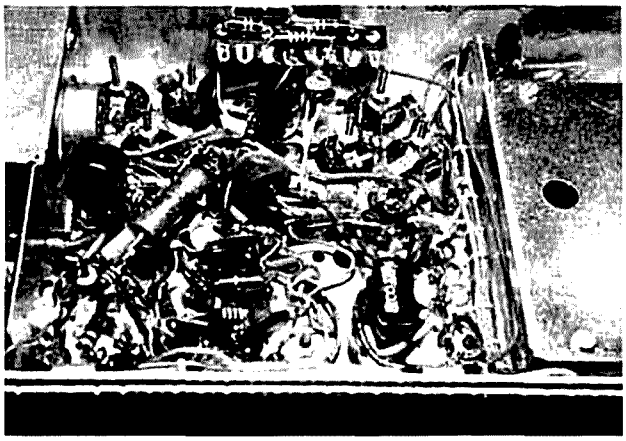


A SOOPAR and power supply mounted on the Permakay filter of a Motorola Sensicon G receiver. The relay is shown mounted on the power-supply chassis.

a complex voltage divider; their values are chosen to present a potential of about -5 volts to the non-inverting input of the op-amp. The values are also chosen to limit the maximum voltage presented to the op-amp to less than the +12-volt op-amp supply voltage.

hysteresis

The advantage of the SOOPAR to turn on and off with a very small input voltage change is a disadvantage when very weak signals are received. The relay will chatter, going on and off rapidly. The tube-type COR avoids this due to the inherent magnetic "hysteresis" in the relay; that is, once the relay is operated, you can reduce the voltage across the relay somewhat and it will remain energized.



A SOOPAR mounted on the side of the chassis of a Motorola Sensicon A receiver.

The relay's pick-up and drop-out voltages are a major factor in the proper operation of a tube-type COR. Resistors R6 and R7 are added to the SOOPAR to introduce electrical hysteresis. The bias presented to the inverting input of the op-amp is raised when voltage is applied to K1 by the op-amp. The amount of this change in potential is determined by the values of R6 and R7.

The net effect of the hysteresis is to cause the relay to release at a SOOPAR input voltage lower than the input voltage required to operate the relay. For example, if the relay drops out at -3 volts (at the non-inverting, or + op-amp input), and picks up at -4 volts, it will release again at -4 volts if there is no hysteresis. By adding hysteresis, it drops out at -3.5 volts.

Incidentally, if you don't want it, this action is called "slop" or "backlash;" it is called "hysteresis" if you do want it. If more hysteresis is desired in the SOOPAR, raise the value of R6 — try 220k or 330k. For less hysteresis try 47k; if no hysteresis is desired, disconnect R7 and short out R6.

receiver connection

Typical connections to Motorola receivers are shown in table 1. Resistors R1 and R4 should be connected directly to the tube socket pins shown, and C1 should be connected directly to R4 and ground. The lead length from R1, R2 and R3 to the op-amp should be kept at a minimum to reduce stray rf pickup. The op-amp itself should be mounted inside the receiver; relay K1 may be mounted externally.

If you adapt the SOOPAR to a different receiver, connect R1, R2, R3, R4 and C1 before connecting the op-amp. Measure the voltage at the junction of R1, R2 and R3 with a vtvm or high-impedance fet volt-meter. It should be between -3 and -6 volts with the receiver set for nominal squelch sensitivity, and should not exceed -12 volts with a saturating signal applied to the receiver input. If necessary, adjust R3 (and perhaps R2) to satisfy these conditions, then

table 1. Connection points for installing the signal-operated operational relay into Motorola fm receivers.

typical Motorola receiver	lead X	lead Y	R3 value
Sensicon A high-band PA 8433/8476	V309 pin 1	V312 pin 5	2.2M
Sensicon G high-band TA 140	V110 pin 1	V113 pin 7	not used
T44 type uhf rcvr TA 141	V14B pin 8	V12B pin 2	2.2M

connect the remainder of the SOOPAR circuit.

adjustment

The threshold adjustment is R5. First adjust the receiver squelch control to quiet the receiver, then adjust R5 to the point where K1 just releases. This adjustment should suffice for most applications, but R5 may be trimmed as necessary to track the SOOPAR operation with the audio squelch.

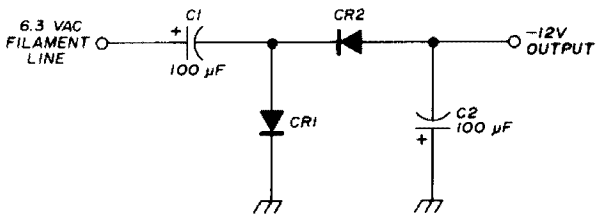


fig. 5. Power supply for the op-amp relay. Diodes CR1 and CR2 are silicon units.

construction

As long as the input leads are kept short, the layout of the SOOPAR is not critical. Small 8-lug terminal strips have been used by some builders. Spare terminals on existing terminal strips within receivers have been used by others. The recommended relay, a Sigma 65FP1-12DC, is a 90-mW, 12-volt dc relay with a 1600-ohm coil. These are available from Allied Radio and others for less than $3.00 each.

A simple voltage-doubler power supply for the SOOPAR is shown in fig. 5. Any small power diodes may be used for CR1 and CR2, and any value of capacitance of 100 µF or more will suffice for C1 and C2. This power supply may be built on a small multi-lug terminal strip similar to the SOOPAR.

conclusion

Several SOOPARs are presently in use in Northern California repeater receivers, including the vhf input receivers of the Mt. Vaca Radio Club's repeaters. This circuit has proven itself reliable under temperature extremes and continuous usage.

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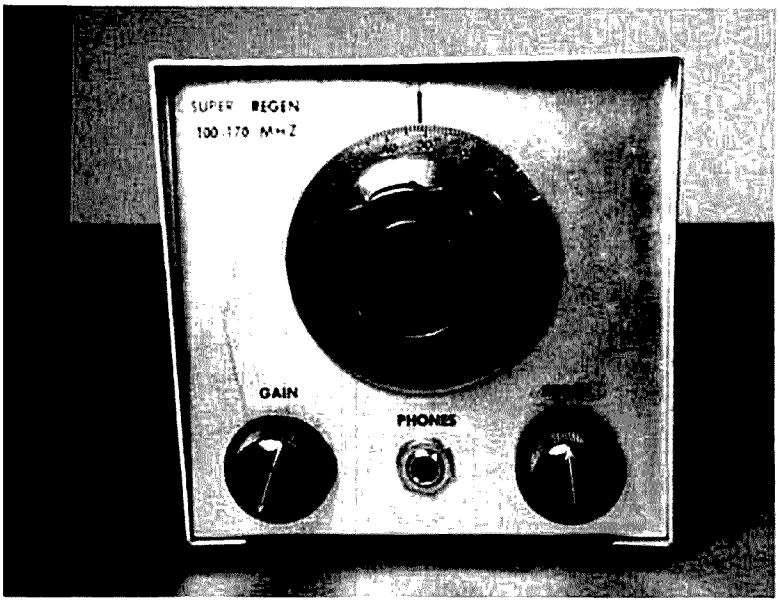
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low-voltage superregenerative receiver for vhf

This superregenerative receiver tunes from 100 to 170 MHz and operates from a single flashlight battery

Courtney Hall, WA5SNZ, 7716 La Verdura Drive, Dallas, Texas 75240

This receiver was built as an experiment to investigate the quality of reception that could be expected from various modulation modes using a superregenerative detector, and to experiment with the diode tank loading technique discussed in previous articles.[1,2] Another design goal was to operate the receiver from a single flashlight battery and to achieve long battery life at low cost. The frequency range of approximately 100 to 170 MHz covers wide-band fm in the fm broadcast band, a-m in the aircraft band and narrow-band fm in the public service band.

Since superregenerative detector theory is covered in the above referenced articles and elsewhere, I will not go into that here.

circuitry

As shown in fig. 1, the receiver consists of a superregenerative detector followed by a three-stage audio amplifier which drives a pair of 2000-ohm headphones. Regeneration is controlled by varying the base bias on the detector transistor. A two-section RC filter between the detector and audio amplifier

attenuates the detector's quench frequency, and each stage of the audio amplifier is "rolled off" with a 180-pF feedback capacitor to prevent response at the quench frequency.

The TIS97 transistors used in the audio amplifier have dc current gains of around 500; if lower beta transistors are substituted, the 150k base-bias resistors

is housed in an LMB 555N aluminum cabinet, which is a five-inch cube. A piece of copper-clad board is held on the inside of the front panel by the tuning capacitor's mounting bushing. This provides a ground plane to which all of the detector's ground connections are soldered.

All detector wiring should be as short and direct as possible. In my receiver

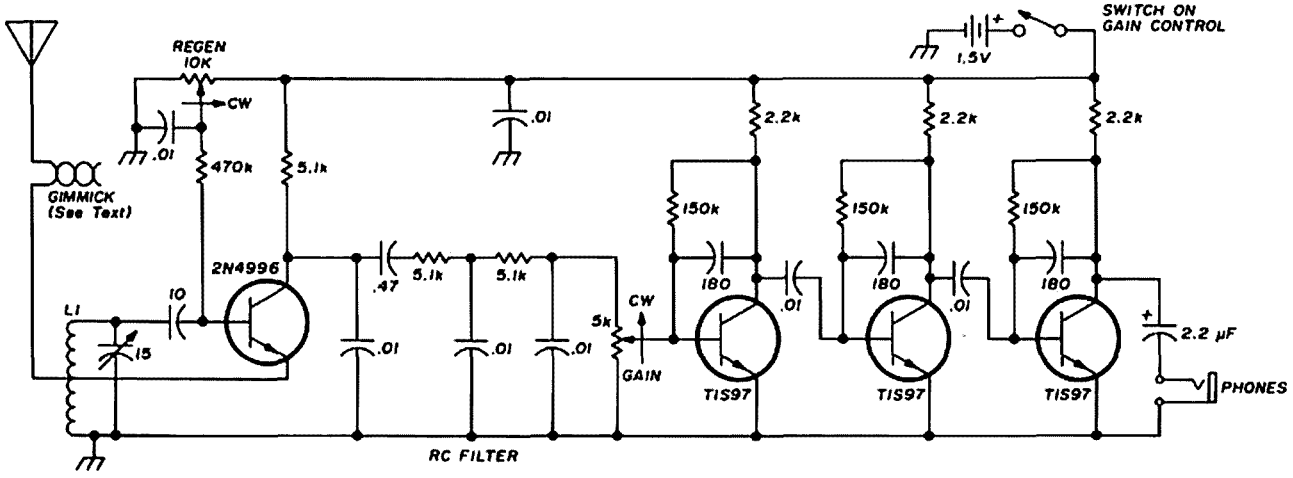


fig. 1. Schematic of the low-voltage superregenerative receiver that tunes from 100 to 170 MHz. Coil L1 is 3 turns Mini-ductor stock, ½" diameter, 8 turns per inch.

may need to be decreased to set the dc collector voltage at about 0.8 volt. Since current drain is about one milliampere, an AA penlight cell could be used instead of the D cell I used.

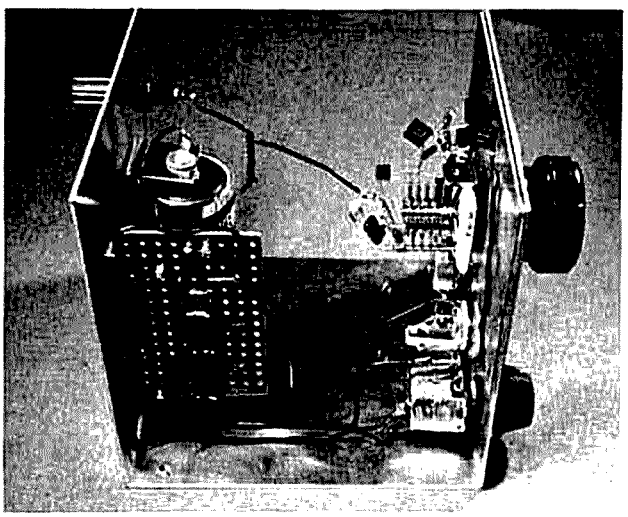
The tuning capacitor is a Hammarlund HF-15-X having about 15-pF maximum capacitance; the coil is three turns of Miniductor stock, 0.5 inch diameter, 8 turns-per-inch. The emitter and antenna are connected to the center of the coil. I found that if the antenna was coupled too tightly to the tank circuit, the detector would not operate properly at the high end of its frequency range.

I used a gimmick capacitor made from two pieces of number-22 hookup wire twisted together for a length of 1.25 inches and inserted between the antenna terminal and the coil. This seemed to provide a good compromise between sensitivity and stable operation when using a three-foot wire antenna.

construction

The entire receiver, including battery,

most of the detector components are just soldered together and supported by their leads. The audio amplifier is assembled on a piece of perf-board mounted on the rear panel with the battery holder and antenna terminal.



Construction of the vhf superregenerative receiver. Receiver components are mounted with short, direct leads around tuning capacitor on front panel. Audio power module is mounted on perf board in rear of enclosure.

operation

For best results, the regeneration control should be advanced to a point just above the spot where the rushing sound begins; this adjustment is not critical, but may need to be varied somewhat with tuning. The receiver should be tuned slightly to one side of an fm station's frequency to achieve slope detection.

Using a three-foot wire antenna, local fm broadcast stations and aircraft came in loud and clear; the aircraft signals are a-m while the fm broadcast signals are wideband fm. Narrowband fm signals in the public-service band were difficult to copy; although their carriers can completely quiet the receiver's noise, their low deviation doesn't provide much audio output from the detector.

I experimented with germanium diodes connected across part or all of the tank coil as described in the referenced articles, but I was unable to detect any improvement in performance.

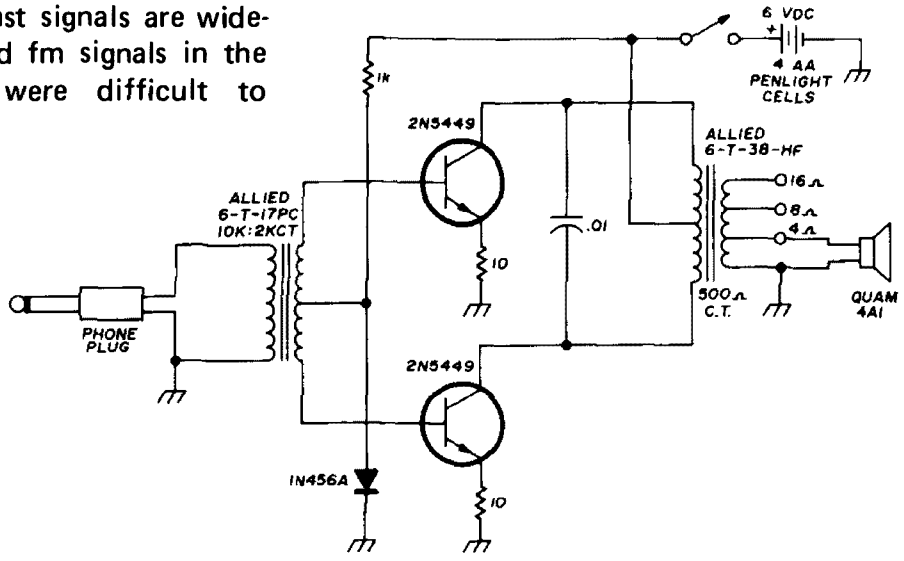


fig. 2. Simple audio power amplifier may be used with any small receiver which is capable of driving a pair of 2000-ohm headphones.

audio power amplifier

Most of my audio circuits are designed to drive 2000-ohm headphones; these phones are easy to drive, and the rest of the household doesn't want to listen to all that shortwave racket, anyway. At times, however, it's desirable to have speaker capability so I built a simple power amplifier which plugs into the phone jack (fig. 2). It is mounted with a 4-inch speaker and batteries in a 6x6x4-inch LMB 664 aluminum box.

Any circuit which will drive a pair of 2000-ohm headphones will drive this unit and produce plenty of room volume. Although I have used only 2N697s and 2N5449s, the circuit should work well with almost any medium power silicon npn devices.

Allied Radio part numbers are shown for the transformers, but any of the less expensive imported transformers having the same impedance levels should be ok. No volume control is included because I usually have one on the equipment designed to drive the phones. Battery drain is 10 mA or less, depending on output level, and four AA penlight cells connected in series will last a long time.

conclusion

Superregenerative receivers can be designed to operate from a 1.5 volt dc supply; they are suitable for a-m and wideband fm reception, but their performance with narrowband fm is poor. Operation of these detectors from a very low voltage source can decrease their radiated power and lessen the possibility of interference to other nearby receivers.

references

1. C.L. Ring, "Optimizing the Superregenerative Detector," *ham radio*, July, 1972, page 32.
2. A. Iwakami, "Improved Superregenerative Receiver," *ham radio*, December, 1970, page 48.

ham radio

importance of standing-wave ratios

This discussion of standing-wave ratios should clear up many of the misunderstandings that persist among many radio amateurs

Earl W. Whyman, W2HB, 375 Mount Vernon Road, Snyder, New York 14226

Twenty five years ago the Micromatch, the first of many vswr bridges designed for use by the radio amateur, was introduced in the pages of *QST*.[1] Prior to that time the radio amateur had not given much attention to vswr and his understanding on the subject was accordingly limited. The fact that vswr measurement might provide a means of effectively evaluating and monitoring the performance of an antenna system was an exciting prospect, and amateurs began to realize that the vswr bridge was a significant development.

Thus, vswr became the subject of many technical articles in the amateur journals, with each writer addressing himself to a discussion of the various phenomenon associated with the presence of standing waves. A considerable bibliography on the subject of vswr has been accumulated, but, nevertheless, a brief period of listening on the amateur bands will reveal that a great deal of misunderstanding continues to exist. For example, it is still a widespread belief that radiated power will increase with decreasing vswr; that reflected power represents lost power; and that the reflected power is somehow dissipated within the transmitting system. One of the invalid theories postulates the reflected power returning to the output stage of the transmitter where it causes overheating of the output tube.

The following discussion is aimed at enabling the reader to obtain a better understanding of the effects of standing waves, and to do so in terms of transmission line theory with which the radio amateur is already well acquainted. In particular, the role of the reflected wave, which has become such an enigma as a result of so many conflicting opinions, will be explained in a way that I believe to be unique.

Part of the discussion will be in the form of a description of experiments performed upon a transmission system which has been mismatched to cause a vswr of 2:1 to exist. In addition, the effects of system vswr upon the output network of the transmitter will be discussed with the aid of numerical examples, and it will be shown that the criteria

of proper operation is simply the ability to tune the transmitter for normal rated plate current.

Knowledge of basic transmission line theory, such as found in the *ARRL Handbook,* is all that is necessary to understand this material. However, a brief review of this theory, with particular emphasis on the aspects most pertinent to this discussion, is included with this article in the form of an appendix. It is suggested that you first review this appendix if only to identify the theoretical aspects which I wish to emphasize.

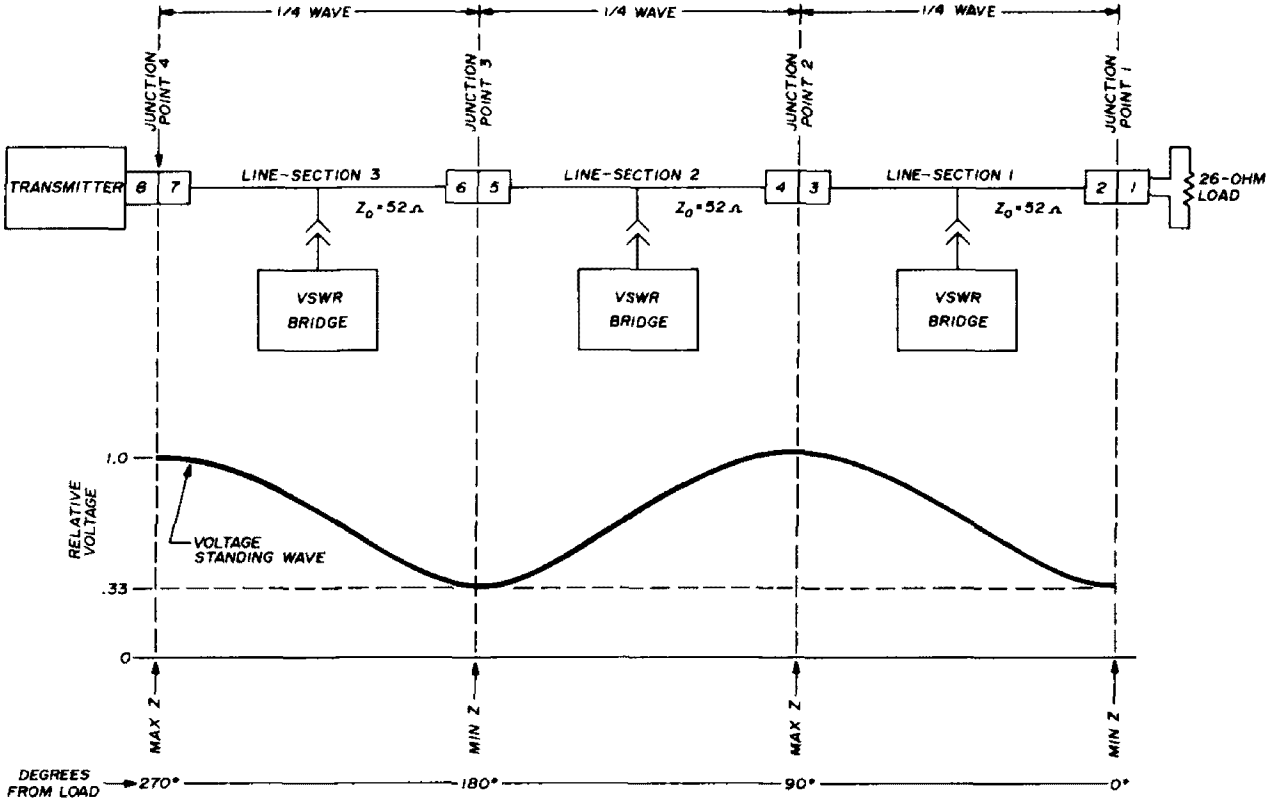


fig. 1. Antenna system discussed in text includes a 3/4-wavelength transmission line with a vswr of 2:1. For diagram of transmitter matching network, see fig. 4.

the experimental system

The most common form of amateur antenna installation in use today uses coaxial cable for the transmission line. Therefore, it is appropriate to base this discussion on the coaxial system shown in fig. 1 where provisions have been made to facilitate the performance of some experimental tests. Connectors have been installed in the transmission line, dividing it into three identical line-sections, each one-quarter wavelength (90°) long. The total length is, therefore, 3/4-wavelength (270°).

The transmission line may be readily opened at the junction points for measurements, and for other purposes to be detailed later in this article. Each line-section, each junction-point and each connector have been assigned a number for reference in the text. In addition, means for monitoring the vswr on each line-section has been provided.

Although a means for opening the transmission line at the resistive impedance points is essential for the experiments to be performed, it is not necessary to control total line length or to select a specific load impedance. The values selected for the system of fig. 1 are simply convenient and lead to less complication in the analytical discussion.

Fig. 1 also shows the position of the

standing wave and the location of the points of maximum and minimum impedance. From transmission line theory we know that the maximum and minimum impedance points are the only points on the transmission line where the impedance is a pure resistance. Later it will be shown that the points of pure resistance are of particular significance in understanding the action of reflected waves.

The system of fig. 1 has other conditions to be specially noted and kept in mind: First, each line-section is made of RG-8A/U coaxial cable which has a characteristic impedance (Z_o) of 52 ohms; second, a vswr of 2:1 exists on each line-section as a result of the 26-ohm resistive load at connector no. 1. Finally, each line-section may be considered separately as a quarter-wave impedance transformer where the impedance at its input connector (connectors 3, 5 and 7) may be readily found by dividing the square of line impedance, Z_o (52^2 = 2704), by the impedance at its output connector (connectors 2, 4 and 6). Thus, the impedance at connector 3 is 2704/26 = 104 ohms. Since connector 3 and connector 4 are joined together, the impedance at connector 5 is 2704/104 - 26 ohms. In a like manner, the impedance at connector 7 is found to be 104 ohms; this is the impedance seen by the transmitter.

experiments and observations

Referring again to fig. 1, let's conduct our first experiment by inserting a short section of transmission line between connector 3 and connector 4 so that the system appears as shown in fig. 2. A section of transmission line with a characteristic impedance of 104 ohms is chosen for this purpose so that it will be perfectly terminated (matched) by the 104-ohm resistive load which we know exists at connector 3.

We also know that when a transmission line is operated in this manner there will be no standing wave on it, and the input impedance will be equal to the characteristic impedance, Z_o, regardless of the length of the line. Therefore, at connector 2A the impedance is 104 ohms, and the load impedance on line-section 2 remains unchanged. Indeed, if you made detailed comparisons of vswr and impedance between the system of fig. 1 and the system of fig. 2, you would be unable to detect any differences — in spite of the fact that the latter system includes the 104-ohm line-section on which there is no reflected wave. The only change is some additional phase delay which is not important to this system.

An interesting and extremely important point has been demonstrated by this experiment. Since no standing wave is present on the 104-ohm line-section, the reflected wave on line-section 1 has ceased to exist at the junction of connectors 3 and 1A. More significant, however, is the fact that the reflected wave has ceased to exist at a point on the transmission line where the impedance is a pure resistance. And, if you were to repeat the above experiment for all other points where the impedance is a pure resistance, the results would be the same.

Thus, in the system of fig. 1 there are three points on the line where the reflected waves cease to exist. However, the impedance at junction-point 2 (104 ohms) is a mismatch for line-section 2, and the impedance at junction-point 3 (26 ohms) is a mismatch for line-section 3. Thus, each of the three line-sections is mismatched by the load impedance at its respective output connector, and the magnitude of the mismatch is such as to cause a vswr of 2:1 on each line-section. Reflected waves are therefore developed independently on each line-section and it is apparent that the reflected wave due to a mismatched antenna load does not travel back toward the transmitter without interruption.

Although it may be somewhat redundant to discuss further the change that was made to convert the system of fig. 1 into the system of fig. 2, some readers may appreciate further clarification of the role of the inserted line-section. Suppose that we experiment further with the

system of fig. 2 by shortening the 104-ohm line-section in successive steps, each time cutting the remaining line-section in half. Theoretically, the line-section would never be reduced to zero length. However, from a practical standpoint you can see that, as a limit, its length would approach zero, thereby making the systems of fig. 1 and fig. 2 identical. Furthermore, if the system is examined

applied years ago in the design of the "Q-Match Antenna."[2,3] In this system the 75-ohm impedance of a center-fed half-wave dipole is matched to 600-ohm open-wire transmission line by using a quarter-wave section of transmission line with a Z_0 of 212 ohms. The 212-ohm line operates with a vswr of 2.8:1, and the 600-ohm line with a vswr of 1:1. Thus, as demonstrated in the foregoing, the re-

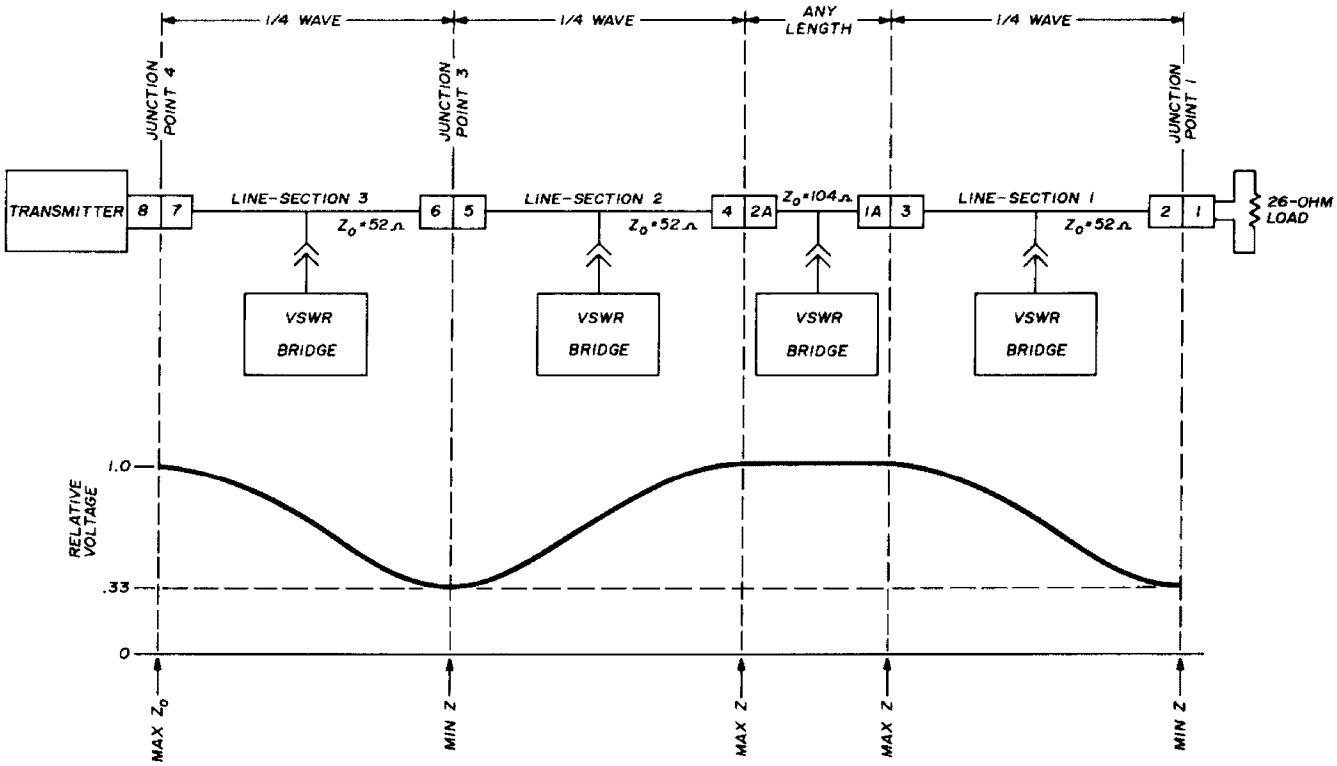


fig. 2. Since the impedance at junction point 2 in fig. 1 is 104 ohms, a section of 104-ohm transmission line, any length, may be inserted at that point with no effect upon the performance of the rest of the transmission line.

after each shortening operation, you would find the operational characteristics (such as vswr and impedance at each junction-point) unchanged. Therefore, it must be concluded that the system is independent of the length of the 104-ohm line-section, including zero length.

The 104-ohm line-section in this case has served as an analytical tool to enable the behavior of the reflected wave to be observed, and to further prove that this behavior remains unchanged after the line-section is removed from the system.

The principle applied in the foregoing experiments is the basis of many impedance-matching systems and was

flected wave on the 212-ohm line ceases to exist at the junction of the two lines because the impedance exhibited at that point is a pure resistance of 600 ohms.

effects on the transmitter

Most present day transmitter designs use a pi-network to couple the antenna system to the output stage. In general the pi networks are designed for a nominal load impedance of 50 ohms at a maximum vswr of 2:1. This means that the transmitter can be loaded to its rated power input as long as the load impedance is maintained within specific limits. By specifying a maximum allowable vswr

the manufacturer is indirectly specifying the impedance limits of the transmitter and the amateur is provided with a parameter he is able to measure.

The easiest way to determine the impedance values which this specification permits is to make a plot on a Smith chart. Simply scribe a circle, with a radius equivalent to a vswr of 2:1, about the center of the chart. This circle is the locus of all possible values of impedance on a system with a vswr of 2:1. A Smith chart with a vswr = 2:1 plot, based upon a Z_o of 52 ohms, is shown in fig. 3; points 1 and 2 on the circle locate the only resistive impedances (26 ohms and 104 ohms, respectively) and point 3 locates the complex impedance of 41.6 + j31.2 ohms for later use in this discussion.

When designing a pi network the value of inductance, L, is determined such that the inductive reactance

$$X_L = \frac{QR_1 + \left(\frac{R_1 R_2}{X_{C2}}\right)}{Q^2 + 1}$$

This expression contains two terms which may be separately equated, one to X_{L1} and the other to X_{L2}, the sum of which is equal to X_L. The significance of X_{L1} and X_{L2} lies in the fact that any pi network can be shown to be exactly equivalent to two L-networks connected back-to-back as shown in fig. 4A. It is no coincidence therefore that X_{L1} and X_{L2} are the respective values of inductive reactance required for L-network 1 and L-network 2. The junction point of the two L-networks must be an impedance match. In fact, if you were to analyze any specific pi-network design you would find that at the junction point the impedance is always a pure resistance with a value lower than either the input or output impedances. For convenience, this point will be referred to as the intermediate impedance of the pi network. Referring again to fig. 4A, capacitors C1 and C2 are the conventional tuning and loading controls, respectively, and L is an inductance value that is fixed for each band. Specific values of reactance for each of these elements is required when a particular pi-network design is desired.

As an example, a pi-network designed to match a single 6146 AB1 linear amplifier to a nominal 52-ohm resistive load is shown in fig. 4B. This network is designed to provide a 3000-ohm resistive load for the 6146 and to operate at a Q of 15. In this case the intermediate impedance is 13.27 ohms. When a load

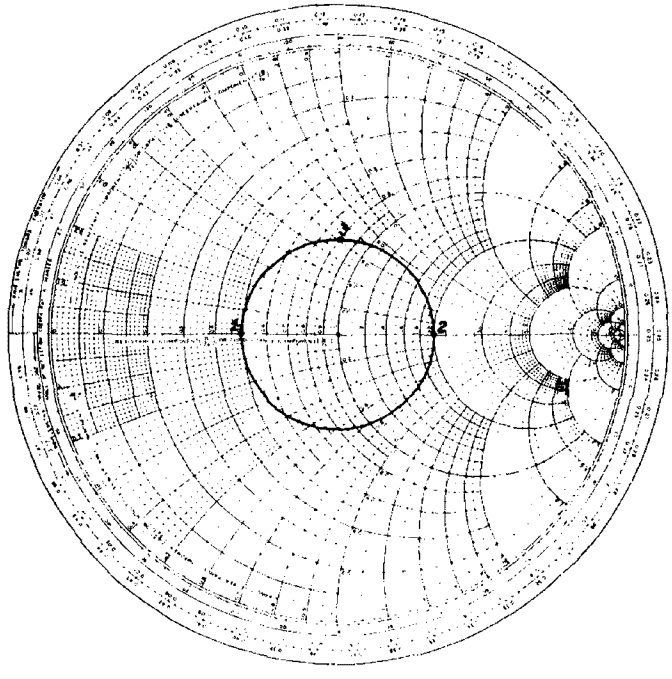


fig. 3. Smith chart plot of vswr of 2:1. Only two pure resistances are possible: point 1 and 2 at 26 and 104 ohms, respectively. Point 3 locates the complex impedance, 41.6 + j31.2 ohms, discussed in the text. (Since this is a normalized Smith chart, all values on the chart must be multiplied by the characteristic impedance of the system, 52 ohms in this case.)

impedance other than 52 ohms resistive is used with this network, C1 and C2 are simply adjusted until normal rated plate current is obtained. When this is done the plate load will always be 3000 ohms resistive — however, the intermediate impedance and the operating Q of the network will change. Fig. 4C shows the values of X_{C1} and X_{C2} readjusted to match a load of 104 ohms resistive; the intermediate impedance is now 11.95 ohms and the operating Q has increased slightly to about 15.8.

When the load at the transmitter's

terminals is complex, i.e., some combination of resistance and reactance, the reactance will affect the tuning of the output network in the transmitter.

Every value of complex impedance can be shown to be simultaneously equivalent to either a resistance in series with a reactance, or a resistance in parallel with a reactance. Thus, there is a choice of using either a series or parallel reactance can be determined for a match to this value of resistive load. This value of X_{C2} is then combined with the required compensating reactance to obtain the final circuit value of X_{C2}.

The result of applying the above procedure to a pi-network designed for use with the 6146 linear amplifier is shown in fig. 5. Fig. 5A illustrates that when an 86-ohm capacitive reactance, X_{CL}, is

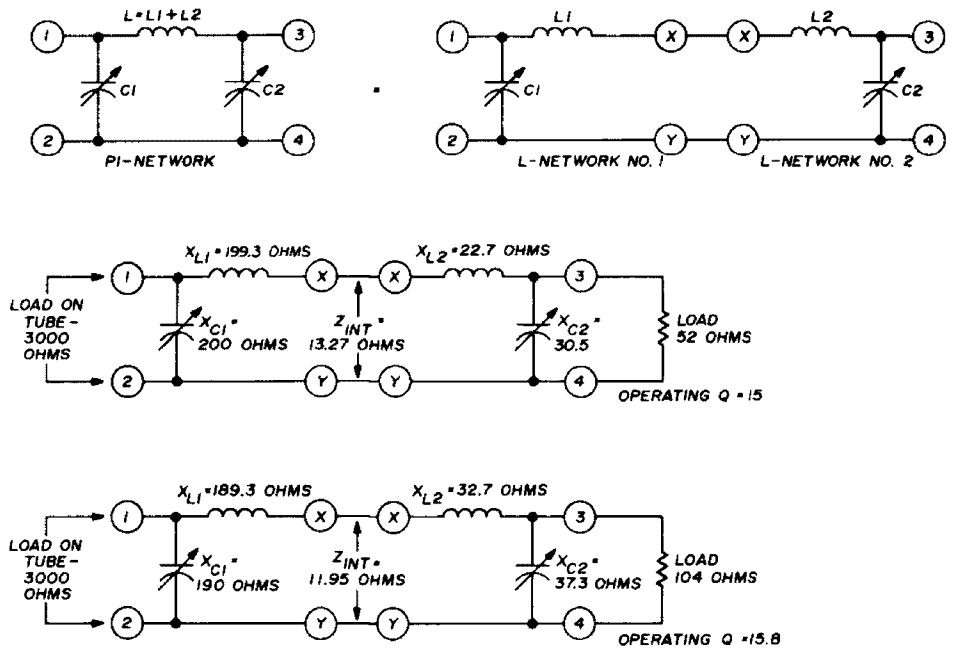


fig. 4. The inductor in the basic pi network consists of two theoretical parts, L1 and L2 (A). When operating into the design load of 52 ohms the operating Q is 15 and the intermediate impedance is 13.27 ohms (B). With a 104-ohm load (vswr = 2:1) the operating Q is 15.8 and the intermediate impedance is 11.95 ohms (C).

to tune out, or compensate for, the load reactance. However, the value of the compensating reactance, and the value of the resulting resistive load, are different in each case.

Thus, point 3 on the Smith chart, fig. 3, is a complex impedance consisting of a 41.6-ohm resistance in series with a 31.2-ohm inductive reactance; its parallel equivalent is a resistance of 65 ohms and an inductive reactance of 86 ohms. The parallel form is of special interest because C2 in the pi network is ideally situated to provide the required compensating reactance in addition to performing its normal function. Since the compensating reactance converts the load impedance into a pure resistance, values of X_{C1} and X_{C2} shunted across the parallel equivalent of the above complex lead the result is a pure resistance of 65 ohms. Fig. 5B shows the network of fig. 4B tuned for a load of 65 ohms resistive. Fig. 5C shows the final network values when X_{C2} is adjusted to include the compensating reactance. Note that the network's internal impedance is 12.6 ohms and its operating Q is 15.6. Thus, the capacitance of C2 had to be increased to supply the compensating reactance. If the load had contained a capacitive reactance, the capacitance of C2 would have to be decreased to achieve compensation.

In practice, the dual role of C2 is obscured by the ease with which the circuit can be adjusted. As long as normal

rated plate current can be obtained the compensation takes place without the operator being aware of it. If it is found that the transmitter cannot be tuned for normal plate current, the vswr is probably in excess of that specified for the transmitter and the output network is unable to provide the required amount of compensation. The solution is either to revise the antenna system or to use some type of antenna tuner.

Figs. **4B**, **4C** and **5C** demonstrate that a pi network designed for a nominal load of 52 ohms resistive can be used with an antenna system on which the vswr is 2:1. In each case the network is tunable to provide a pure resistance load of 3000 ohms for the plate circuit of the tube, the internal impedance of the pi network is a pure resistance, and the deviation from the desired operating Q is minimal.

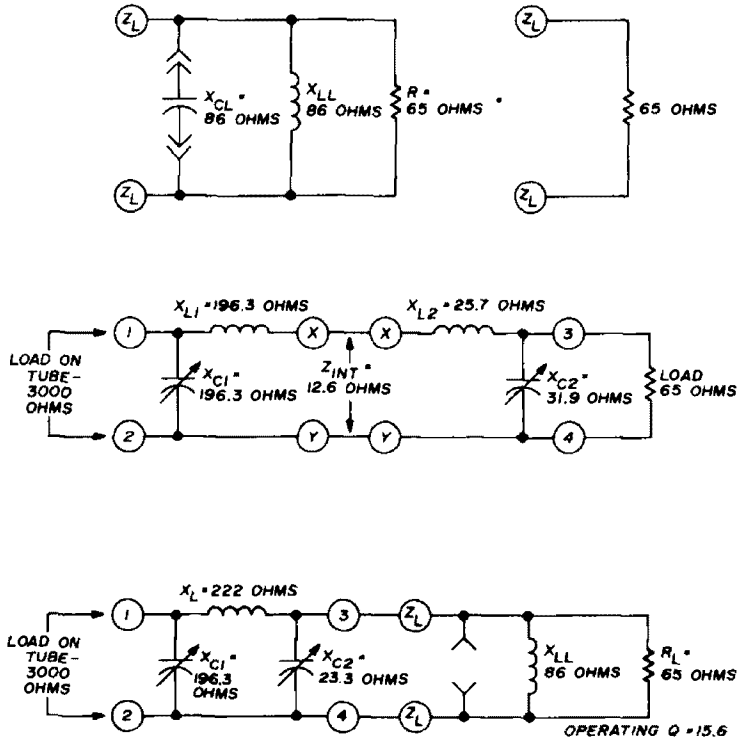


fig. 5. A capacitive reactance may be shunted across an inductive load to tune out the inductance, leaving a resistance of 65 ohms (A). The pi network of fig. 4 may be adjusted to match a resistive 65-ohm load (B). Final pi network reactance values, when C2 has been adjusted to include the compensating capacitive reactance, are shown in (C).

summary

Throughout this article the points of resistive impedance have been given special emphasis. When the experiments performed on the transmission line were discussed, the points of resistive impedance were significant because it was shown that reflected waves cease at these points. When the operation of the pi network was discussed it was shown that, for moderate values of system vswr, the network can always be tuned to produce the proper value of resistive load for the power tube and the intermediate impedance is always a pure resistance.

Whenever a pure resistance exists in a circuit, at that point power can flow in only one direction. Therefore, power must flow from the generator (or transmitter) toward, and into, the point of pure resistance. This is so because resistance is strictly passive. That is, it cannot reflect, and all of the power that flows into it must be absorbed. A resistive point in a circuit is no different than any other resistor in this respect. Therefore, you should not expect it to act any differently when it occurs in a transmission line system.

Thus, you can see that there is no way for power to return to the transmitter, and except for minor transmission line losses, the transmitter's power has no place to go except into the antenna, regardless of the presence of standing waves.

appendix

Standing waves on a transmission line are the result of a reflected voltage wave combining with (or interferring with) the forward voltage wave. Since the two

waves are traveling in opposite directions at a constant velocity, the resulting rms voltage at any point on the transmission line will be constant and will depend upon the relative phase and amplitude of the two waves.

Locations of peak voltage will occur where the forward and reverse wave voltages are in exact phase and add together; in the same manner, voltage nulls will occur where the forward wave voltage and the reverse wave voltage are in exact opposite phase and the resulting amplitude is the difference between the two voltages. Peaks and nulls will be alternately spaced at one-quarter wave intervals (90 electrical degrees) along the transmission line with their number dependent upon the electrical length of the line.

The vswr is obviously the ratio of the voltage at peak to the voltage at a null and may be stated algebraically as

$$\text{vswr} = \frac{E_f + E_r}{E_f - E_r} \qquad (1)$$

where E_f is the voltage amplitude of the forward wave and E_r is the voltage amplitude of the reflected wave. Eq. 1 can be rearranged, and if at the same time E_f is assigned a value of unity, a new expression is obtained

$$E_r = \frac{\text{vswr} - 1}{\text{vswr} + 1} \qquad (2)$$

The amplitude of the reflected wave voltage can now be expressed as a fraction of the forward wave voltage for any value of vswr. For example, if the vswr is 2:1, the reflected wave will be 1/3 the amplitude of the forward wave.

Transmission-line theory is frequently discussed in what is termed "the lossless case." The assumption is made that there is no loss in the dielectric material of the transmission line and no losses in the electrical conductors. Nevertheless, in the practical case losses do exist in both the dielectric and the conductors. However, if the system being considered is limited in length to a few wavelengths, and if the operating frequency is not above the high-frequency range, there is little error involved in assuming the losses to be zero and the discussion becomes much more manageable.

If the transmission-line losses are assumed to be zero, the product of line voltage and line current must be the same for any point on the line. It follows that to satisfy this condition of constant power, a standing wave of current must exist such that the current nulls coincide with the voltage peaks, and the current peaks coincide with the voltage nulls. Since impedance is the ratio of voltage to current, $Z = E/I$, it should be noted that high line impedance occurs at voltage peaks, and low line impedance occurs at voltage nulls. With a little further manipulation of the foregoing analysis the ratio of maximum impedance, Z_{max}, to minimum impedance, Z_{min}, may be expressed as a function of vswr

$$\frac{Z_{max}}{Z_{min}} = \text{vswr}^2 \qquad (3)$$

This ratio is 4 when the vswr is 2:1.

One quarter-wavelength (90°) sections of transmission line have the unique property of operating as impedance transformers when terminated in any impedance other than Z_o. Thus, the input impedance (Z_{in}) for any value of output impedance (Z_{out}) is readily found from

$$Z_{in} = \frac{Z_o^{\,2}}{Z_{out}} \qquad (4)$$

It is obvious that when Z_{out} is equal to Z_o the input impedance will be equal to Z_o and no impedance transformation will take place. It is also apparent that the input impedance will be a pure resistance only when the output impedance is a pure resistance.

references

1. Jones and Sontheimer, "The Micromatch," *QST*, April, 1947.
2. Marshall, "Antenna Matching with Line Segments," *QST*, September, 1948.
3. Geiser, "Resistive Impedance Matching with Quarter-Wave Lines," *QST*, February, 1964.

ham radio

frequency synthesizer for two-meter fm

Bill Craig, WB4FPK, Post Office Box 947, Grayson, Kentucky 41143

Although this all-channel two-meter frequency synthesizer was designed for the GE Progress Line, it may be used with other equipment

One of the first things a newcomer to fm learns is that he never seems to have enough room in the rig for all the crystals he wants, or needs, especially when using obsolete commercial equipment. While he could go out and buy a commercial amateur transceiver, with multi-channel capabilities, there is a considerable price differential between these and surplus commercial equipment, not to mention that commercial equipment usually has a far better intermodulation figure than most of the amateur rigs currently available. This can be a very valuable asset in areas served by several repeaters that are not very far apart in frequency.

Obviously, to invest in a large number of crystals would be a financial hardship to the ham with a limited pocketbook, and hardly anyone would be interested in purchasing 100-plus crystals just to be able to cover all the available channel combinations, simplex and duplex. My aim in designing this unit was to have a low-cost unit with limited capability. This unit will only synthesize frequencies that are spaced at 30-kHz intervals starting at 146.01 MHz, which allows full coverage of the entire portion of the two-meter band which is available for repeater operation.

Naturally, this presents certain limitations, as in some areas there are still repeaters which are not on "standard" frequencies. But they are (fortunately) few and far between. In a situation like this, the synthesizer could be used to drive one channel of a two-frequency rig, and the special crystals could be installed in the other channel.

Synthesizers have been covered in quite great detail in recent months in various publications, so I refer you to the articles mentioned in the reference list at the conclusion of this article.

While I designed this unit for use with the GE Progress Line equipment, it can be modified to operate with other commercial equipment such as Motorola, RCA, etc. I would be glad to assist anyone interested in adapting the unit to other equipment if they will send me full details of what they want to do, as well as the multiplication factors of the receiver and transmitter and the receiver first i-f frequency, etc., along with a stamped, self-addressed envelope.

To permit rapid channel switching, channel selection is accomplished by means of a 24-position, 10-deck rotary switch, in conjunction with small diode matrices, which are used to convert the binary-coded decimal (BCD) numbers required by the programmable dividers into their decimal values. While this involves considerable wiring at the switch, it also makes operation much simpler, as changing channels only requires a twist of the wrist, as would be compared to some similar circuits which would require setting 6 to 8 thumb-wheel switches. Again, some versatility is sacrificed, but there is a way out of this which I will discuss later.

The 10-deck switch is in reality, two 5-deck switches on a common shaft. A second *band* switch, a 6 pdt rotary switch, selects either one or the other bank of five decks, depending upon whether the frequency desired is above or below 147.00 MHz. In this manner, 24 possible combinations are available in either bank, each covering a full MHz. A dual set of channel markings is provided on the front panel, one set for Band A, the other for Band B. You could simply eliminate all frequencies above 147.00 MHz, and use a 5-deck switch (which would also eliminate the other band-switch) as you would have to retune the transmitter at least in order to operate over the frequency spread that the synthesizer is capable of. The receiver will operate reasonably well over a wider range than the manufacturer specifies, however. A base-station operator might wish to retain the full coverage so he can take advantage of possible DX openings.

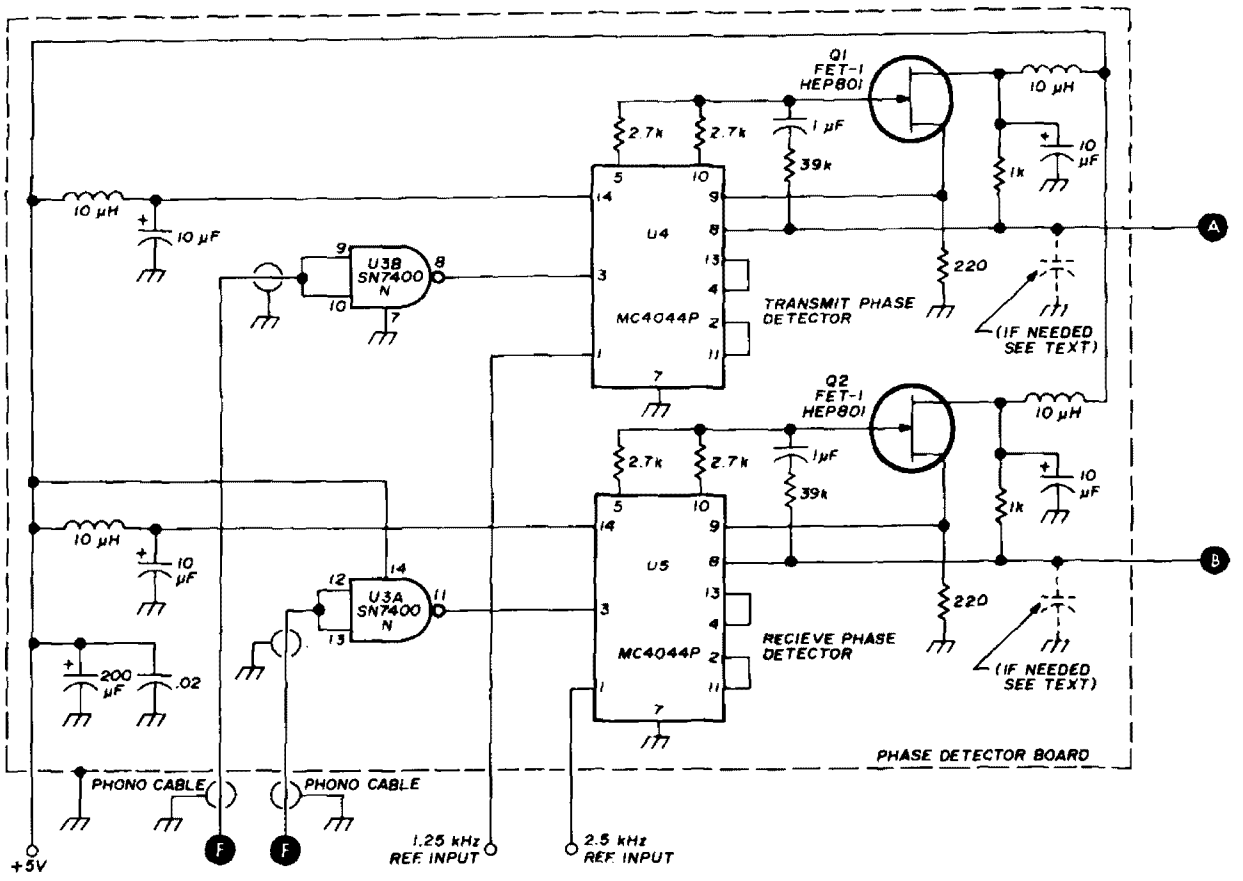
Whenever possible, I used low-cost general-purpose replacement type transistors to eliminate any problem of procuring replacements (these items can be obtained in most any radio/tv supply house in the country). All ICs used were of the lowest cost types available at this time, and can be obtained from most of the specialty houses dealing in these items at very nominal cost.

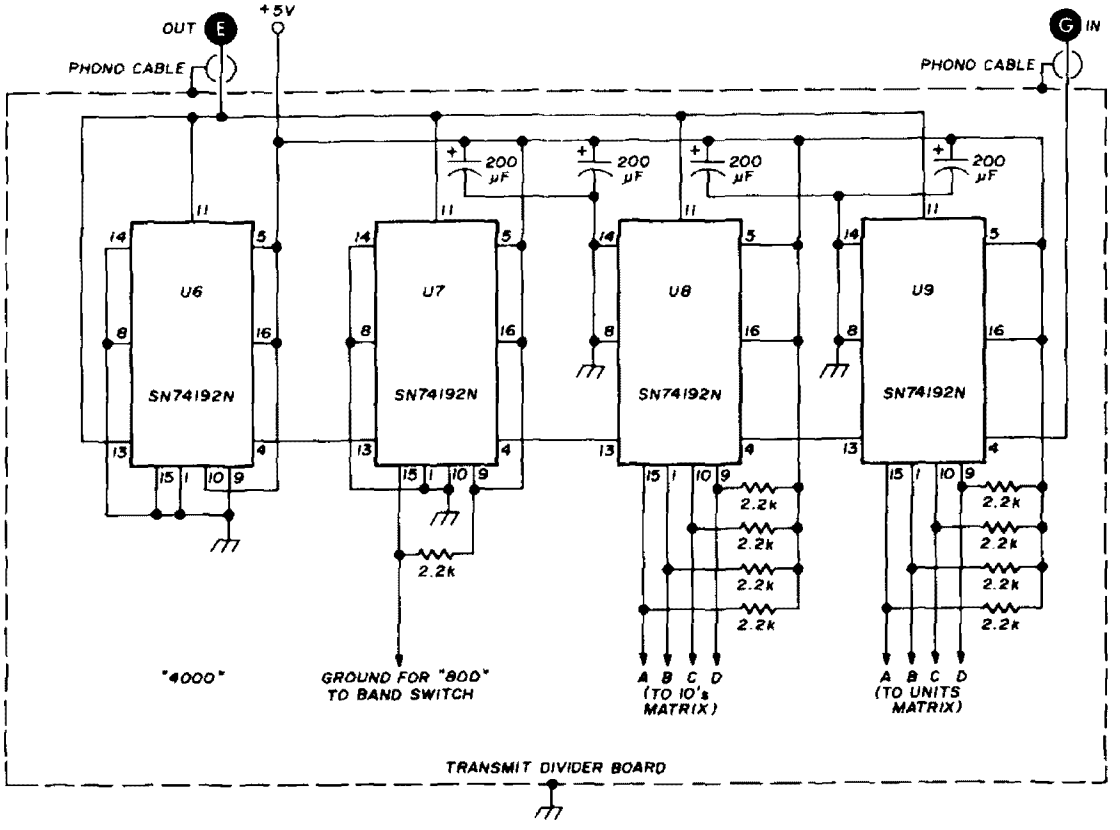
circuit description

The programmable dividers are made from SN74192N decade up/down counters, wired in a modulo n configuration. These divider ICs have a parallel load capability, which allows you to enter a certain count, and then count down to zero, whereupon, an output will be generated and the number will again be reentered into the dividers.

The number is entered in BCD form, as was mentioned previously, which is derived from diode matrices. Note that only the units and 10s of the numbers to be entered require the matrices, as the 100s only change at infrequent points during the coverage of the 2-MHz segment. The 1000s never change and are coded directly at the PC board. The coding of the 100s occurs at two decks of the channel switch.

For those readers not acquainted with BCD numbers, each of the dividers is provided with a parallel input lettered A, B, C and D which correspond to the BDC numbers 1, 2, 4 and 8, respectively. If we were to connect the A input to a 1 voltage (more than approximately +2.0 V), the divider would preset to the count of 1; if we wanted to encode the number 7, we would connect the A, B and C inputs to 1, etc. All other inputs are connected to zero (less than 0.8 V) or grounded. Of course, we would also have to apply the proper signal to the load control lead from this package before we could enter





C1	10 pF, N750
C2	20 pF, N750
CR1,CR2	22-pF varicap (TRW PC-116)
L1	24 turns no. 24 enamelled on Millen 69041 slug-tuned form
L2	10 turns no. 24 enamelled on Millen 69041 slug-tuned form
L3,L4	turns as required on Millen 69041 slug-tuned form (see text)
C_A,C_B	temperature-compensating capacitors (see text)
C_T	0,8 to 11 pF piston capacitor (JFD NVC9G)
X1	5-MHz crystal (International Crystal HA-1)

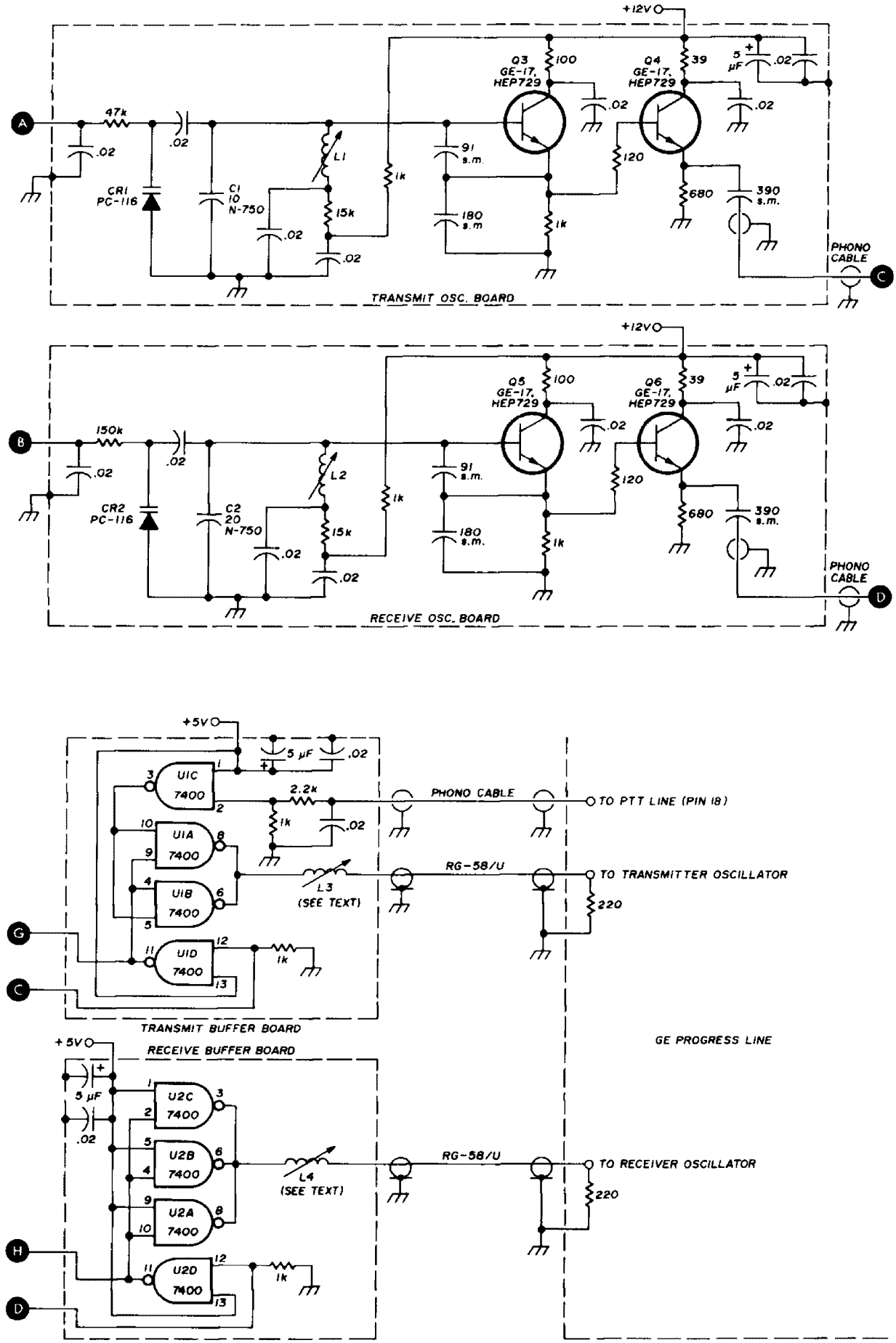


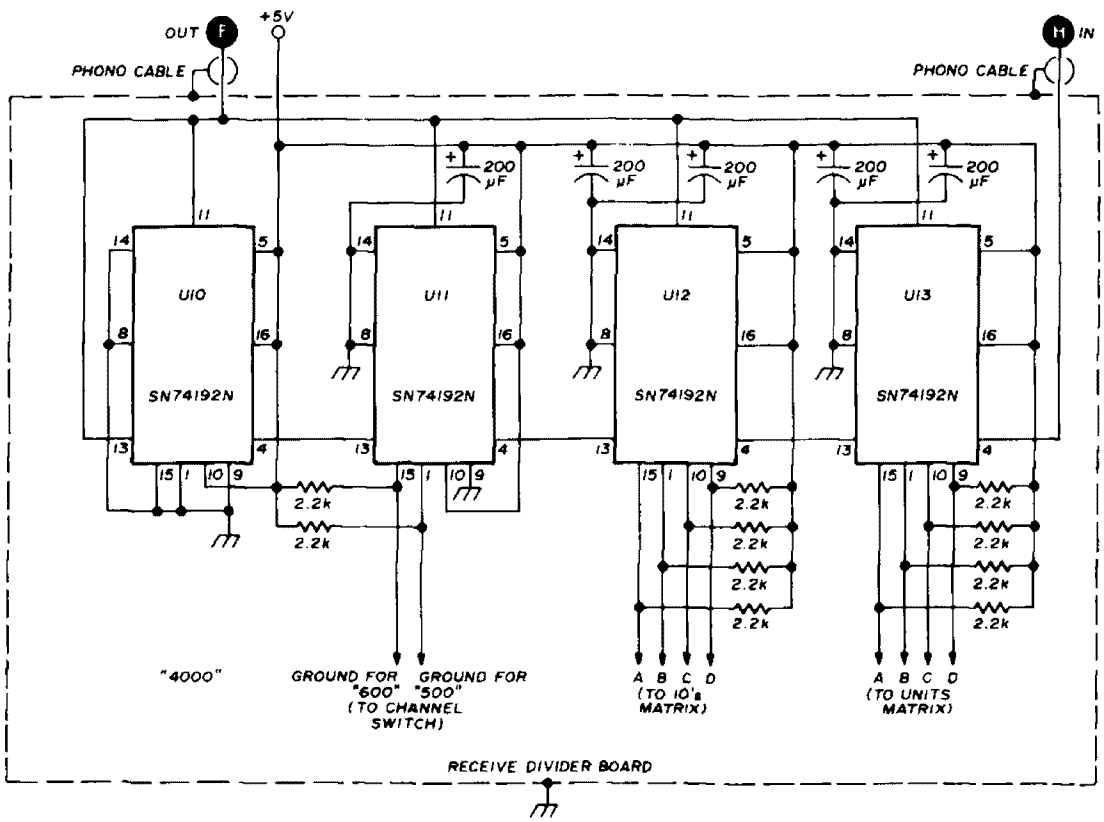
fig. 1. Circuit diagram for the two-meter transmit and receive synthesizer. This unit covers the entire two-meter fm band in 30-kHz steps, starting at 146.01 MHz. The 10-µH rf chokes are J.W. Miller type 9340-20. L1-4 are James Millen type 69041. L1 is 24 turns no. 24 enameled wire. L2 is 10 turns no. 24 enameled wire (scramble wound). Schematic for receiver divider board is on the next page.

the numbers. The load lead requires a zero to enter the data.

Each package has two clock inputs and two outputs marked count up, count down, carry and borrow. In this particular application we are only going to count down. The rules for using this IC require that the unused clock input be connected to 1 voltage. The carry output is not connected to anything in this particular application.

when an output occurs at its borrow, toggling the second divider in the chain one time. The first divider continues to count down 9,8,7, etc., until zero is reached again, and the second divider is again toggled. As you will note, the divider is working as a decade divider after it counts down the initial 6 of our number 4866.

This process continues until the second divider has counted down from



Schematic for the receiver divider board. Circuit points F and H are connected to the circuit in fig. 1, shown on the preceding pages.

Note that all the load inputs are connected to the *borrow* of the last (most significant number) divider. It happens that whenever an output occurs from the *borrow* the output changes from 1 to zero. It becomes apparent then that every time an output occurs from the divider chain the same signal is applied to all the parallel load leads re-entering the number into the dividers.

Suppose we take a short-cut and say we already have the number 4866 entered in our dividers. Now we apply clock pulses to the first count down input and the divider counts down, "6,5,4,3,2,1,0," the initial 60 it was coded with, indicating we have counted the 66 part of our original number, where the second divider also now acts as a decade divider in conjunction with the first divider to divide by 100 which will toggle the third divider in exactly the same fashion until the total number 4866 is reached, whereupon an output will occur at the borrow of the last divider in the chain. This causes the number to be re-entered into all the dividers again though the action of the output signal also being applied to the load leads of each package.

However, one little problem slips in

here, but it's quite simple to solve. Note that the dividers don't toggle on the negative transition of the clock pulse or the borrow pulse, but on the positive transition, meaning that the total count is short-changed by almost a full clock pulse. Note that as soon as the borrow occurs (not at the end of the borrow as is occuring between the sections of the dividers) the new number is re-entered into the dividers again. As soon as this number is reentered, the output obviously cannot any longer be zero because the number 4 has just been entered. So instead of getting an output pulse that is equal in length to one cycle of the input frequency (about 166 nanoseconds at 6 MHz) the output pulse is shortened to something on the order of 5-10 nanoseconds — the switching time of the TTL logic, and counting starts immediately instead of at the end of 166 nanoseconds as would be the case if you were using the more expensive modulo n dividers such as are manufactured by Motorola.

The solution to this problem is quite simple. We merely say that, instead of counting by 4866, we actually counted by 4867. The net result is that the frequency error is quite minute if we simply deduct one number from the division ratios that we calculate (the tables accompanying this article take this into account), and we end up with modulo-n counters at about one-half the price of devices intended specifically for this purpose.

Note that all the inputs that will require programming are connected to 1 via resistors. Those inputs that never change are connected to either zero or 1 directly at the PC board, as is appropriate in each instance. By shorting the inputs that go to 1 to ground (via resistors) through paths developed within the diode matrices or the bandswitch and channel switch, you can program the dividers for any count, within certain limits. The count can be programmed for considerably further range than is listed in the tables, however, if desired.

Since the basic crystal frequency of the GE Progress Line transmitter is in the 6-MHz range, and is multiplied by 24 to the output frequency, you can determine the required division ratio for the divider by dividing the desired output frequency by 24, and then by the reference frequency (1250 Hz in the case of the transmitter) remembering to subtract one count from the answer as explained above. After working out a few of these it becomes apparent that the reference frequency is related to the channel spacing (1.25 kHz x 24 = 30 kHz).

The same procedure is used to determine the division ratio required for the receiver divider, bearing in mind that the multiplication factor is only 12 in this case and subtracting the first i-f frequency from the desired receive frequency before dividing by 12. This figure is then divided by the reference frequency which in the case of the receiver is now 2.5 kHz (2.5 kHz x 12 = 30 kHz). Don't forget to deduct the one count from the final answer here, also.

The output of each programmable divider is buffered through a TTL gate to eliminate any possible transients that may appear in the output signal which are below the threshold level of the gate but which might affect the action of the phase detector. Large value capacitors are provided liberally on the PC boards to help reduce transient amplitude which could alter the total count.

phase detectors

Phase detector circuits are provided for each of the programmable dividers, and their associated vcos. External components must be added to provide an active low-pass filter circuit which is tailored to the referency frequency, the maximum number of divider steps (loop sensitivity), the maximum lockup time desired, and the maximum permissible overshoot. To go into the math for the design of these filters would be beyond the scope of this article and the interested reader is referred to Motorola Application Note AN-535. The use of a fet transistor

rather than a bipolar type in the low-pass filter circuit makes the filtering easier to accomplish as the time constants of the associated R-C components are not affected as badly by the relatively high gate impedance of the fet.

Notice that the circuit provides for rf decoupling in the supply leads to both the active filters and phase detectors. This is required to prevent the steep wave front of the reference and output signals from the dividers from pulse modulating the resulting dc control voltage which is being obtained from the phase detector circuit and which would also modulate the vcos in turn.

voltage controlled oscillators

The vcos are conventional Colpitts design, with a single transistor buffer stage. Frequency variation is accomplished by applying the voltage derived from the phase detectors to a varicap diode wired into the tank circuit. Isolation of the rf is adequately provided by a series resistor to this diode, which also serves the purpose of limiting the swing of the oscillator, as the diode will be somewhat self-biased by some rectified rf from the oscillator tank circuit. This requires that the applied control voltage be raised above this value before any appreciable change in capacitance can occur.

The chosen L/C ratios for the two oscillators cover the required frequency spread by about a ratio of 2:1, ensuring that adequate margin will exist in the event that the oscillator changes frequency due to changes in temperature, but when corrected by the output from the phase detector, will still be within the range of that circuit. Some frequency stability is imparted by the parallel padding capacitors in each tank circuit. However, this could probably be improved somewhat if the builder has available N-1500 type capacitors instead of the specified N-750 types. The oscillators operate from 12 volts, and are ac coupled to the following logic buffer stages. The two oscillators are identical in construction, with the exception of the

table 1. Division ratios for standard repeater pairs and simplex pairs. (Channel 22 of group A, and channels 20, 21, 22, 23 and 24 of group B are spares.)

switch pos	xmit freq (MHz)	receive freq (MHz)	xmit xtal frequency (kHz)	xmit divisor	rcv xtal frequency (kHz)	receive divisor
group A						
1	146.01	146.61	6083.75	4866	11492.5	4596
2	146.04	146.64	6085.00	4867	11495.0	4597
3	146.07	146.67	6086.24	4868	11497.5	4598
4	146.10	146.70	6087.50	4869	11500.0	4599
5	146.13	146.73	6088.75	4870	11502.5	4600
6	146.16	146.76	6090.00	4871	11505.0	4601
7	146.19	146.79	6091.25	4872	11507.5	4602
a	146.22	146.82	6092.50	4873	11510.0	4603
9	146.25	146.85	6093.75	4874	11512.5	4604
10	146.28	146.88	6095.00	4875	11515.0	4605
11	146.31	146.91	6096.25	4876	11517.5	4606
12	146.34	146.94	6097.50	4877	11520.0	4607
13	146.37	146.97	6098.75	4878	11522.5	4608
14	146.40	146.40	6100.00	4879	11475.0	4589
15	146.43	146.43	6101.25	4880	11477.5	4590
16	146.46	146.46	6102.50	4881	11480.0	4591
17	146.49	146.49	6103.75	4882	11482.5	4592
18	146.52	146.52	6105.00	4883	11485.0	4593
19	146.55	146.55	6106.25	4884	11487.5	4594
20	146.58	146.58	6107.50	4885	11490.0	4595
21	146.70	146.70	6112.50	4889	11500.0	4599
22	-	-	-	-	-	-
23	146.94	146.94	6122.50	4897	11520.0	4607
24	147.00	147.60	6125.00	4899	11575.0	4629
group B						
1	147.03	147.63	6126.25	4900	11577.5	4630
2	147.06	147.66	6127.50	4901	11580.0	4631
3	147.09	147.69	6128.75	4902	11582.5	4632
4	147.12	147.72	6130.00	4903	11585.0	4633
5	147.15	147.75	6131.25	4904	11587.5	4634
6	147.18	147.78	6132.50	4905	11590.0	4635
7	147.21	147.81	6133.75	4906	11592.5	4636
8	147.24	147.84	6135.00	4907	11595.0	4637
9	147.27	147.87	6136.25	4908	11597.5	4638
10	147.30	147.90	6137.50	4909	11600.0	4639
11	147.33	147.93	6138.75	4910	11602.5	4640
12	147.36	147.96	6140.00	4911	11605.0	4641
13	147.39	147.99	6141.25	4912	11607.5	4642
14	147.42	147.42	6142.50	4913	11560.0	4623
15	147.45	147.45	6143.75	4914	11562.5	4624
16	147.48	147.48	6145.00	4915	11565.0	4625
17	147.51	147.51	6146.25	4916	11567.5	4626
18	147.54	147.54	6147.50	4917	11570.0	4627
19	147.57	147.57	6148.75	4918	11572.5	4628

changed circuit values of the tank components and the series resistor to the varicap diode.

logic buffer stages

A quad two-input gate is used to provide direct coupling to the GE unit and the logic divider stage. One section is used as a buffer between the vco and the dividers, and also drives the inputs of two gates wired in parallel that are used as the output buffers to drive the GE unit. Note that by ac coupling the output of the oscillator stage in the synthesizer to this

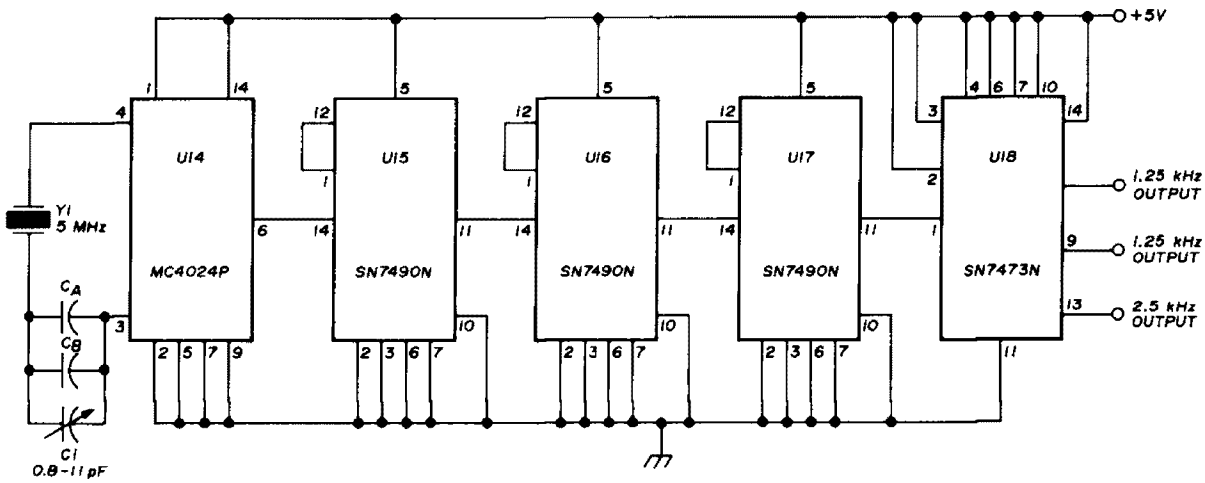


fig. 2. Crystal oscillator reference generator board.

gate, the gate can only respond to that portion of the sine wave output of the oscillator that is 0.8 volt above the zero axis of the ac signal. Consequently, the rise and fall times of the output of the gate are more than adequately fast enough to drive the divider stages without the need of a Schmitt trigger or other pulse-forming device (approximately 40 nanoseconds rise/fall time as measured on a Tektronix 513D scope).

The output from the paralleled gates is fed to the GE unit through a length of coaxial cable. The length is not especially critical, as I tried lengths up to 15 feet and still obtained adequate drive for the multiplier stages, even though the end of this line was terminated in a 220-ohm resistor.

The two buffer stages are essentially alike in their function, although the receiver buffer has three gates tied in parallel, and the transmitter buffer only two, as I wanted to use one gate section as an inverter. Note that the unused inputs to the gates in the transmitter buffer are not returned to +5 volts as they are in the receiver buffer, but are connected to the output of this inverter stage. The input of this inverter is connected to the line running between the transmitter push-to-talk relay and the microphone switch contacts. When the push-to-talk button is released, approximately +12 volts* appears on this line, driving the output of this gate to zero, which disables the two buffer gates. During transmitter operation this same point is at ground and consequently the output of the inverter is at 1, which enables the buffer gates. Since it only controls the buffers and doesn't interrupt the path between the output of the vco and the dividers, no frequency searching occurs when going from transmit to receive. This suppresses the output of the transmitter oscillator signal to a level below the internal noise level of the receiver strip so as not to interfere with weak-signal reception.

This was ascertained with a narrow-

*Some GE Progress Line equipment uses the -22 volt bias supply to operate the relays. In this case connect the center pin to the cathode of the 6AQ5 audio output tube in the receiver instead of connecting it to pin 18 of the Jones connector.

band receiver strip that was carefully checked for sensitivity and found to be somewhat better than the original specifications for this strip. However, the addition of a preamplifier would probably change the entire picture, and it would probably become necessary to supply transmit excitation through a coaxial relay keyed along with the transmitter relays. Good shielding will help in this instance more than any other technique.

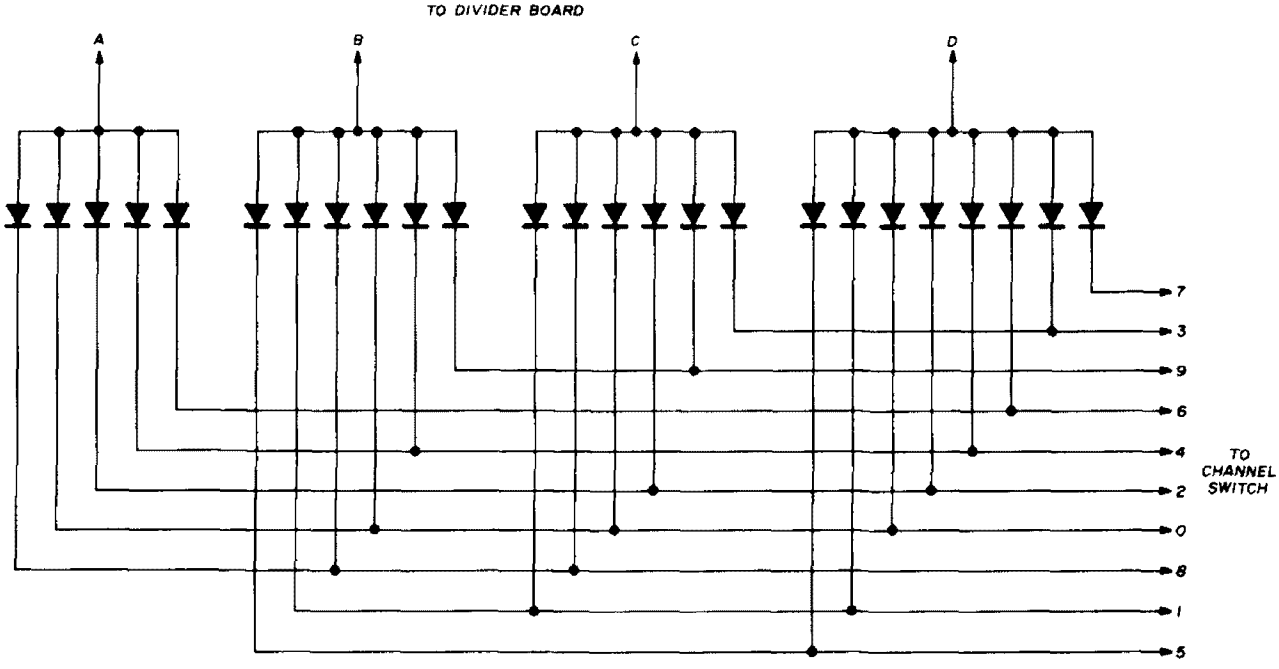


fig. 3. One-half of each diode matrix. All diodes are germanium types with a forward current rating of 10 mA, PiV = 10 volts (minimum).

Of course, this presents no problems when operating duplex as the synthesized transmit frequency is considerably removed from the receiver frequency. If only this type of operation is contemplated the additional circuitry can be deleted. I originally tried tank circuits at each end of the feed line in an attempt to improve the waveform and consequently, reduce the harmonic content. However, the additional complexity of the circuit was considerable as additional drive requirements would have to be provided to make up for losses in all these additional inductors. I feel that the simplicity of this circuit far outweighs any disadvantage it may incur. The small inductors wired in *series between the end of the coax and* the buffers provide a better impedance match to the logic and also act as a rudimentary low-pass filter.

construction details

The printed-circuit boards are stacked to provide as compact a package as possible. With the arrangement shown, all the ICs can be replaced without the necessity of dismantling anything. It is most important that the two oscillators and their two buffers be mounted in as rf tight a package as you can possibly make! Any coupling between the two oscillators will result in spurious outputs due to the fact that the mixing of the two signals will produce sidebands at intervals of 312.5 and 412.5 kHz when operating on repeater pairs spaced at 600 kHz. These spurious outputs will be treated by the transmitter as modulation components due to their equal spacing from the synthesized frequency, and will be multiplied, and appear in the output at the same spacing!

Bear in mind that an fm transmitter, although it multiplies the carrier and the sideband frequencies an identical amount, also causes the modulation index to increase with each multiplication. Consequently, the spurious outputs still

appear at the output. However, these that are apparent are the offspring of the original pairs since multiplication causes additional side frequencies to appear which will be spaced at the same intervals from the carrier as the original pairs were.

The compartment for the two oscillators was made from sheet copper bent into a channel shape to form the sides, and has an internal divider also made of sheet copper, and top and bottom made of PC board. Input power leads, vco control leads, etc., are bypassed immediately inside the compartment. Each oscillator has its own separate power leads. This is also most important to prevent mixing of the two rf signals. The buffer stages are treated in the same manner. However, in this case I constructed the shielded compartments entirely from copper clad PC board, soldering the sides and bottom entirely, and then lining the openings with Eimac finger stock. A rectangular shaped piece was then formed from a strip of ¼-inch wide copper strap, which was made a tight press fit into the opening.

The edges of these pieces were then soldered to lids made from PC board to complete the installation. The circuit boards that are installed inside these compartments are soldered to the sides of the compartment using small no. 6 solder lugs bent at right angles to reinforce the solder joint. Leads are then run from the boards to the appropriate terminal connections on the wall of the enclosure.

This type of construction is most necessary at these points to prevent rf from the output of the transmitter from being re-amplified which would cause hum, excessively wide bandwidth of the transmitter output, audio distortion, etc. The series inductors that I mentioned earlier in the discussion of the buffer stages also help to some extent to decouple the rf path that would otherwise exist on the center conductor of the coax. BNC connectors were used extensively here, which also helps in this regard.

Good quality phono connectors such as the Switchcraft 3502 plugs and 3501FP jacks are a good choice for the reference leads going from the buffer compartments to the programmable dividers. These plugs are completely shielded, and the jacks only require a single ¼-inch hole for mounting.

Notice that the 24-position channel switch is mounted directly to the two matrice boards (front two boards) that contain the four diode matrices. This provides for very short leads here (not really a necessity here as only dc paths are located here) which makes for a neater layout. The front board contains the two matrices for encoding the units of the *receive and transmit* dividers, the board immediately to the rear of this contains the two matrices for encoding the receive and transmit 10s of the dividers, *and the* board next to the rear of that position is the transmit divider board. To the rear of that board is the receive divider board, the phase-detector board, then the compartments housing the two oscillators, and lastly, the compartments housing the buffers.

The diodes used in the matrices are germanium types to permit remote location of the channel switch if desired, as this would allow for almost 0.5 V drop across the interconnecting wiring. This type of diode is recommended in all cases to provide margin for error, and to allow future additions to the equipment. This would permit the synthesizer to be trunk mounted in a mobile installation with just the channel switch mounted at the dash in a Minibox.

Not included in my rig are the timebase and power supply boards, which were not used with this particular unit as the reference and supply voltages were obtained from a frequency standard that was already in the shack. However, I have included the necessary information for building these boards. The crystal oscillator board and its dividers can be mounted to the left of the stack of boards so the trimmer capacitor would be accessible from the front panel through a small opening. The power supply board can be

mounted on the righthand side or wherever is most convenient to the constructor.

The oscillator board uses the Motorola dual-vco package (only one-half used) as the crystal oscillator circuit. This makes construction quite simple as the only external components required are the crystal, trimmer capacitor and a temperature compensating capacitor. A high quality crystal is recommended here, as the output frequency stability of the synthesizer can never be any better than the stability of the reference frequency. The International type HA-1 crystal is recommended for this purpose. It should be ordered for room temperature, 12-pF parallel load for this circuit (HC6 type holder).

For mobile use an oven and suitable oven-type crystal would be required, especially in northern climates. The oscillator is then followed by three decade dividers, whose output is divided by two by the first section of a dual J-K flip-flop, yielding the 2.5-kHz reference output. This is then again divided by two by the other half of the flip-flop producing the 1.25-kHz reference output.

Due to variations in crystal and component tolerances and characteristics, the determination of the fixed temperature compensating capacitor is left to the builder. As a start, set the trimmer to approximately half capacity and install a 10-pF N-750 capacitor on the pads provided (CA). Tune in WWV (at the highest frequency possible) and attempt to zero beat the oscillator to WWV's signal by adjusting the trimmer. It may be necessary to add or subtract some capacitance from CA in order to get the oscillator in the proper range of the trimmer. Once in zero beat, hang a 100-watt lamp a foot or so above the board and allow the board to be warmed by this for about an hour.

Unless you are extremely lucky, the oscillator should no longer be in zero beat with WWV. Carefully noting which way you have to turn it, adjust the trimmer to bring the oscillator back into zero beat again. If you had to turn it clockwise (more capacitance) then the oscillator is over-compensated. Remove the capacitor from CA and install one of half that value (still type N-750) and in addition, install another capacitor of the same value at CB (type NPO) and repeat the test. If, after the initial test, you instead had to turn the capacitor counter-clockwise (less capacitance) the oscillator was not sufficient-

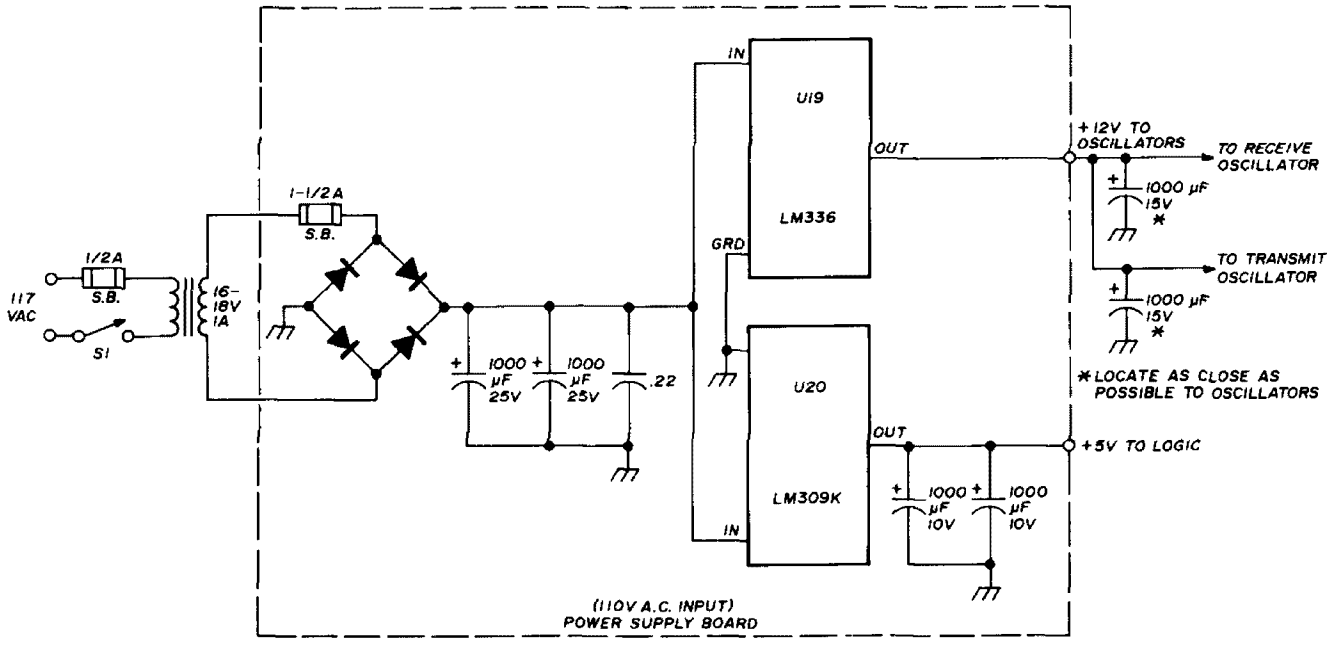


fig. 4. Ac power supply for the frequency synthesizer. Transformer T1 is a control or doorbell transformer with 16 to 18 volts output across the secondary.

ly compensated. In this case install a N-1500 type capacitor as a replacement for the original N-750 type and again repeat the test. (The circuit board should be permitted to cool between tests so as not to give meaningless indications.) By using various ratios of capacitance at CA and CB (the former being either N-750 or N-1500, the latter NPO) any intermediate value of compensation can be obtained.

power supply

The power supply is quite simple, using two self-contained, protected ICs to provide the regulated outputs. The power transformer can be a doorbell transformer or control transformer with 16-18 volts output at 1 amp. This output is rectified by a bridge circuit and applied to the two regulators. A fuse is provided in the secondary (as well as the primary) in the unlikely event of a short in one of the rectifiers or ICs. The 5-volt regulator is a National Semiconductor LM-309K (in a TO-3 case). The 12-volt regulator can be either the National LM-336 (in TO-5 case) or the Fairchild μA7812 (TO-220 case). For mobile use, the 12-volt regulator is deleted, the 12-volt battery is applied directly to the 5-volt regulator input, and also to a simple shunt zener diode regulator operating at 10-volts. (This will require readjustment of the oscillators with the lower supply voltage, however.) The output from the zener regulator is additionally filtered through four 88-mH toroids wired in series.

If PC board layouts shown* are used, the construction of the boards is just a matter of mounting parts. It is strongly recommended that you use IC sockets on all the boards (as well as transistor sockets in the two oscillator boards) to make changing of defective ICs and transistors simple, as some of the PC board wiring is quite minute and could easily be damaged by repeated soldering and unsoldering of components. Also, in the oscillator compartments, the sockets would eliminate the need of unsoldering the small printed-circuit boards from the enclosure if a transistor becomes defective. After all the boards are mounted, solder a heavy braid from each board to the next with as short a lead as possible.

All cables that are indicated as shielded on the schematic diagram should be RG-58/U coax unless otherwise indicated. Both ends of the cable shield must be grounded to the nearest ground point and the inner conductor should be as little exposed as possible. Note that all power and control leads leaving or entering the vco and buffer shielded compartments must be bypassed immediately within the compartment with 0.02-μF disc capacitors with the shortest possible leads, soldered directly to the copper surface of the compartment. (The small electrolytic capacitors indicated are also mounted in a similar fashion.) The power and control leads were made through miniature pin jacks to make it easy to disconnect a sub-assembly for repair or testing. Feedthrough capacitors could also be used in lieu of these but would require that the leads be unsoldered to disconnect it from the rest of the wiring.

synthesizer adjustment

After completing construction, connect the logic circuits to the output of the 5-volt supply with a milliammeter inserted in series (0-1 amp range). Note the current, which should be between 450-550 mA for the logic circuits, exclusive of the crystal oscillator and divider board. If it is more or less than this range you have either a defective IC or filter capacitor or something that isn't connected. If this checks ok, connect the oscillators to the 12-volt supply with a series milliammeter (0-100 mA scale) and note the current here, which should be 30-40 mA (both oscillators). If all checks well to this point, connect a voltmeter (20,000 ohm/volt or vtvm) from the vco control line to the receiver oscillator (and to ground) and turn the channel switch to

*Full-size templates of the original prototype printed-circuit boards are available from the author along with additional schematics and parts placement diagrams for $1.00 postpaid.

the highest receive frequency (refer to the chart, as the highest receive frequency is not necessarily the highest switch position) *and adjust* the slug *in* the receive oscillator coil form (L2) for a reading of 3.8-4.0 V (use the 0-10 volt scale). Then turn the channel switch to the lowest receive frequency and note the meter reading, which should be between 2.5-3.0 volts. If outside these limits, the L/C ratio

exterior wall of the compartment and the threaded shank should be used to make the adjustments.

troubleshooting

If the voltage stays at or near zero or +4.8 volts, some problem exists in the loop circuits. This can be caused by any number of interrelated problems such as no reference input to the phase detector,

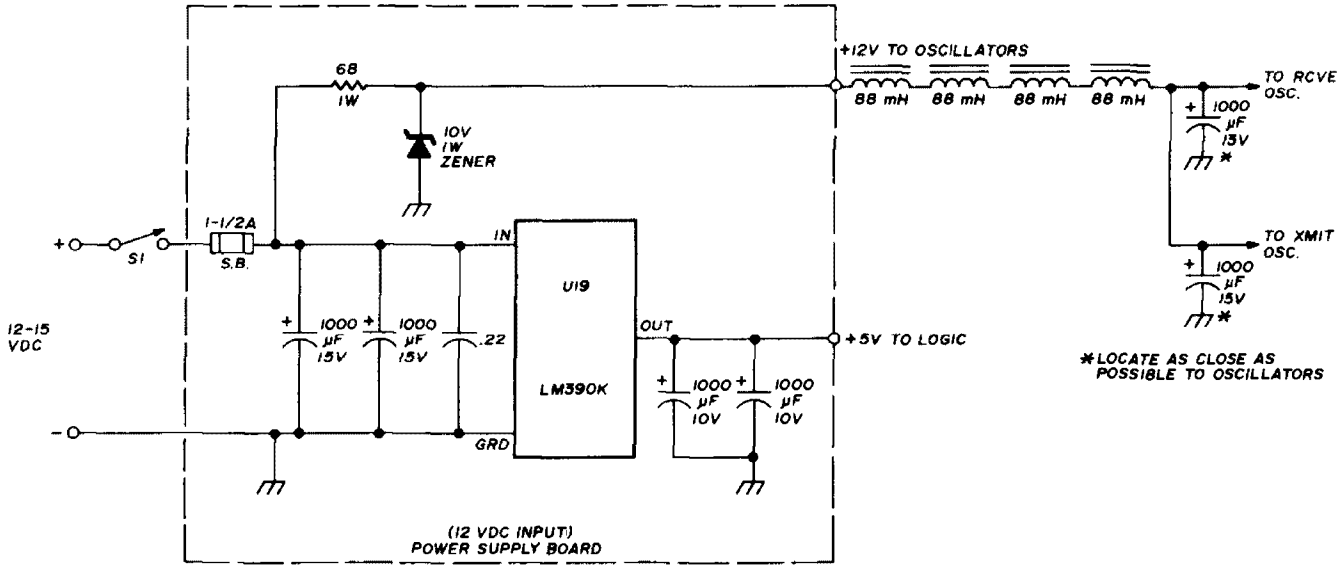


fig. 5. Alternate dc power supply for 12 to 15 Vdc input.

of the oscillator tank circuit must be adjusted, since this indicates that the oscillator control sensitivity is not correct.

If the voltage is too high, add more parallel padder capacitance to the tank circuit and reduce total inductance. Repeat the adjustment procedure until the voltage falls within these limits. If the voltage is too low, remove some of the parallel padder capacitance and increase the inductance.

Repeat this procedure for the transmit oscillator (L1), checking at the transmit vco control lead. This is probably locking the barn door after the horse has strayed, but I hope you read all the instructions before starting on this project. The oscillator compartments should be provided with access holes for adjustment of the coil slugs. Do not remove the compartment covers to make this adjustment. The coil forms should not be mounted to the

no input to or output from the programmable dividers, improper operation of oscillators or buffers, or oscillator frequency so far removed from the desired frequency that the loop cannot make the necessary corrections. Obviously, some troubleshooting will be required here and a good oscilloscope will be quite useful.

Check to see if the oscillators are actually operating near the desired frequencies by listening for their outputs on a general-coverage receiver. The oscillators should, with the vco control lead disconnected, be operating about 200 kHz below the lowest frequency in the list. Generally speaking, a reading of +4.8 volts on the vco control lead indicates that there is no input from the programmable dividers to the phase detector (or no input to the dividers from the buffer stage, etc.) or the oscillator is tuned too low in frequency.

A reading of zero (or nearly zero)

indicates that there is no input to the phase detector from the reference frequency source, or that the oscillator is tuned too high in frequency. A general-coverage receiver is handy at this point to determine if there is any output from the oscillators at all or any output from the buffer stages.

Once you have determined that the oscillators are working and the loops are locking, tune in the output frequencies on a general-coverage receiver and, with the bfo turned on, note whether the tone is stable and reasonably pure. (Don't overload the frontend of the receiver as this will give erroneous results. Keep the rf gain control turned as low as possible while making this test.)

table 2. Channel switch connections for programming per table 2. Column heading AT indicates group A, transmit; BT: group B, transmit; AR: group A, receive; BR: group B, receive.

switch pos	Deck 10 units AT	Deck 9 units BT	Deck 8 units AR	Deck 7 units BR	Deck 6 10s AT	Deck 5 10s BT	Deck 4 10s AR	Deck 3 10s BR	Deck 2 100s AR	Deck 1 100s BR
1	6	0	6	0	6	0	9	3	5	6
2	7	1	7	1	6	0	9	3	5	6
3	8	2	8	2	6	0	9	3	5	6
4	9	3	9	3	6	0	9	3	5	6
5	0	4	0	4	7	0	0	3	6	6
6	1	5	1	5	7	0	0	3	6	6
7	2	6	2	6	7	0	0	3	6	6
8	3	7	3	7	7	0	0	3	6	6
9	4	8	4	8	7	0	0	3	6	6
10	5	9	5	9	7	0	0	3	6	6
11	6	0	6	0	7	1	0	4	6	6
12	7	1	7	1	7	1	0	4	6	6
13	8	2	8	2	7	1	0	4	6	6
14	9	3	9	3	7	1	8	2	5	6
15	0	4	0	4	8	1	9	2	5	6
16	1	5	1	5	8	1	9	2	5	6
17	2	6	2	6	8	1	9	2	5	6
18	3	7	3	7	8	1	9	2	5	6
19	4	8	4	8	8	1	9	2	5	6
20	5	spare	5	spare	8	spare	9	spare	5	spare
21	9	spare	9	spare	8	spare	9	spare	5	spare
22	spare	spare	spare	spare	spare	spare	spare	spare	spare	spare
23	7	spare	7	spare	9	spare	0	spare	6	spare
24	9	spare	9	spare	9	spare	2	spare	6	spare

interconnections

Once you have determined that the synthesizer is working correctly, the next step is to make the necessary connections to the GE unit. Begin by installing a phono jack on the front panel as near to the receiver oscillator crystal socket as possible. Attach one of the ends of a short piece (6 to 8 inches) of RG-58/U coax to this jack, and install a suitable plug on the other end that will mate with the crystal socket. (A suitable plug can be made from a defunct FT-241 crystal holder. Some radio/tv supply houses carry a twin-lead plug that also has the correct dimensions.) Solder a 220-ohm resistor across the plug, too.

The grounded side of the plug should go in the crystal pin hole that is nearest the right-hand edge of the receiver chassis (when looking from the front). Install a known, good crystal and measure the multiplier voltage at the jack provided on

the chassis with a 20,000 ohm/volt meter set on the 0-2.5 V scale. Note this reading. Remove the crystal and connect the plug from the synthesizer instead, and again note the reading. It should be nearly the same or higher.

Adjust the series inductor in the buffer compartment (L4) to obtain the highest possible reading and tune the synthesizer to a channel that is in use. Observe if the audio output from the receiver is distortion free and if there is anything more than just the slightest trace of the 2.5-kHz reference frequency apparent. (If everything is working correctly you should only be able to hear, and then just barely, the reference frequency with receiver gain wide open when listening to a strong unmodulated carrier.)

If an objectionable level of reference frequency is apparent it is probably due to insufficient shielding of the oscillator in the synthesizer, or of the buffer stages, or insufficient bypassing of the supply and control leads. It is possible, but not very likely, that the components used in the phase detector are incorrect in value. However, there is quite a bit of leeway here, and the values would have to be drastically different from those specified to cause this. In some instances, adjustment of the series inductor in the buffer compartment will reduce the amplitude of the reference-frequency trace and should be tried first. However, after the final adjustment you should still have adequate meter indication at the multiplier test jack.

Assuming that the receiver is now working correctly, install a BNC connector on the front panel of the GE unit as far to the left as possible from the jack installed for the receiver. This should be a bulkhead-type connector so that the cable shielding is uninterrupted. Attach an 18-inch length of RG-58/U cable to this connector and install a phono plug on the other end. Route this cable down the left-hand edge of the receiver chassis, around the rear of the receiver to the left-hand corner of the transmitter chassis. There is just enough room in the left rear corner of the transmitter chassis to mount a phono jack. From the center pin of this jack, run a short stiff wire to the grid of the A oscillator tube (single-channel versions) or a short length of miniature coax to the grid of the B oscillator tube (two-frequency versions). Wire a 220-ohm resistor across the existing 100k grid resistor. For best results it is strongly recommended that you convert the oscillator tube to an amplifier by bypassing the screen and cathode of the tube with .02-μF disc capacitors.

If you only have a single-channel rig it is suggested that you convert it to a two-frequency version for this application. You don't need any special parts such as trimmers or temperature compensating capacitors if it's only going to be used as an amplifier, and this will leave the A channel available for crystal operation if so desired. In the two-frequency version, the screen lead of the B oscillator should be disconnected from the screen lead of the A channel oscillator tube socket, and then connected through a 47k resistor to the B+ point. This will prevent the added bypassing at the screen and cathode of the B oscillator from bypassing the A oscillator as well. Refer to the appropriate GE manual for your rig for the proper connections for making a two-frequency rig from your unit. It's a simple job.

tuneup

The transmitter can now be connected to a dummy load, placed in the tune position, and, with a multimeter connected to the first multiplier test jack, note the voltage which is obtained from a known, good crystal. (If you have a single frequency model, make this measurement before converting the oscillator to an amplifier and note the reading for future reference.) Connect the synthesizer, preferably at or near the same frequency as the crystal was, and compare the readings. These should be as nearly alike as possible, and may be changed by adjusting the series inductor in the buffer compartment (L3).

If the voltage is too high, you probably have too much series inductance; too low, not sufficient inductance. About 8 to 10 turns of no. 24 enameled wire is correct for a three-foot length of cable between the synthesizer and the front panel of the GE unit. This is approximately correct for the receiver coil, too, by the way. With a 15-foot length of cable, about 18 turns is correct. As a rough starting point, the inductive reactance of the coil should be approximately equal to the capacitive reactance of the cable. The voltage at the multiplier test point must remain nearly the same as with a crystal as otherwise the modulation index will change.

table 3. Division ratios for all frequencies between 147.99 and 146.01 MHz.

frequency (MHz)	transmit divisor	receive divisor	frequency (MHz)	transmit divisor	receive divisor
146.01	4866	4576	147.00	4899	4609
146.04	4867	4577	147.03	4900	4610
146.07	4868	4578	147.06	4901	4611
146.10	4869	4579	147.09	4902	4612
146.13	4870	4580	147.12	4903	4613
146.16	4871	4581	147.15	4904	4614
146.19	4872	4582	147.18	4905	4615
146.22	4873	4583	147.21	4906	4616
146.25	4874	4584	147.24	4907	4617
146.28	4875	4585	147.27	4908	4618
146.31	4876	4586	147.30	4909	4619
146.34	4877	4587	147.33	4910	4620
146.37	4878	4588	147.36	4911	4621
146.40	4879	4589	147.39	4912	4622
146.43	4880	4590	147.42	4913	4623
146.46	4881	4591	147.45	4914	4624
146.49	4882	4592	147.48	4915	4625
146.52	4883	4593	147.51	4916	4626
146.55	4884	4594	147.54	4917	4627
146.58	4885	4595	147.57	4918	4628
146.61	4886	4596	147.60	4919	4629
146.64	4887	4597	147.63	4920	4630
146.67	4888	4598	147.66	4921	4631
146.70	4889	4599	147.69	4922	4632
146.73	4890	4600	147.72	4923	4633
146.76	4891	4601	147.75	4924	4634
146.79	4892	4602	147.78	4925	4635
146.82	4893	4603	147.81	4926	4636
146.85	4894	4604	147.84	4927	4637
146.88	4895	4605	147.87	4928	4638
146.91	4896	4606	147.90	4929	4639
146.94	4897	4607	147.93	4930	4640
146.97	4898	4608	147.96	4931	4641
147.00	4899	4609	147.99	4932	4642

Next, listen to the transmitter on a good fm receiver and carefully adjust the series coil, until you null out any hum that is apparent on the carrier. Hum in this case is usually caused by rf feedback, and careful adjustment will usually eliminate the path through the coax. If the hum persists, and the adjustment of the inductor doesn't seem to change it, then possibly rf is getting back into the buffer or oscillator stages by *another* path. All the shielding should be carefully checked to eliminate any other entry points for rf. If this eliminates the hum switch to high power and listen again. Note if there is any audio distortion or echo apparent; this is also an indication of rf feedback.

If everything is fine up to this point, listen carefully to the unmodulated carrier with an fm receiver for noise (this noise, if present, will probably sound like weak static). If this condition is noted, try adding a 5-μF capacitor from the transmit vco control line to ground, and increase this value up to 20-μF to eliminate this condition. This problem is caused by variation in tolerances in the components used in the low-pass filter in the phase detector. Don't add this capacitor unless it's really needed, as it will slow down the loop lock-up time.

At this point the only remaining thing to do is to install a phono jack on the front panel of the GE unit near the Jones connector going to the control head. From the center pin of this jack run a lead to pin no. 18 of the Jones connector. Install a shielded cable from the jack to the control connection to the inverter stage in the transmitter buffer compartment. This circuit will suppress the output from the transmit buffer during receiving periods.

The oscillator board (crystal oscillator) should now be zero beat against WWV. If a frequency counter is available the output of the synthesizer should be meas-

ured and the crystal trimmer adjusted accordingly. It's not a bad idea to check all the output frequencies (including the receiver) to make sure that you actually wired the band and channel switches correctly as well as to make certain that the synthesizer is locking on all frequencies.

other frequencies

As I mentioned earlier there is a way to externally program the unit for unusual channel combinations if desired: for example, you might want a .34 simplex combination for some testing purpose. This is quite simple to do if you leave the last position of the rotary channel switch in the second group (B) of frequencies blank and mark this *external* on the front panel. You then bring out all the decimal leads from the matrices, the hundreds leads from the receiver divider board, and the hundred lead from the transmit divider board. The Units and 10s leads are wired to thumb-wheel switches (the decimal encoding types, not the BCD type) with the hundreds leads from the receiver divider board going to a spdt toggle switch. The common leads from all these switches are returned to the position 24 segments of the B group decks on the channel switch.

The hundreds lead from the transmit divider can be obtained from the unused connection on the bandswitch, run to one side of a spst toggle switch the other side of which is connected to ground.

The proper division ratios are then read from a frequency chart and entered into the switches to obtain these special frequencies. You must remember to leave the transmit hundreds switch in the 900 position, however, whenever using the regular channel selector switch in the B group. (When using the A group this precaution is not necessary as the external switch will be disconnected by the bandswitch and the lead is normally grounded by the bandswitch.) A chart of the division ratios that would have to be set on the thumbwheels and toggle switches is included for frequencies from 146.01 to 147.99 MHz (see table 3).

acknowledgement

I would like to express my thanks to Don Rees, W4VQA, and Chuck de Santis, WB8FQY, who were most helpful in providing suggestions and assistance in the initial testing of the synthesizer, as well as to all the patient people who use the Huntington *machine* who had to live through the birth pains and countless hours of discussion and testing that went on. Their patience and comments are greatly appreciated.

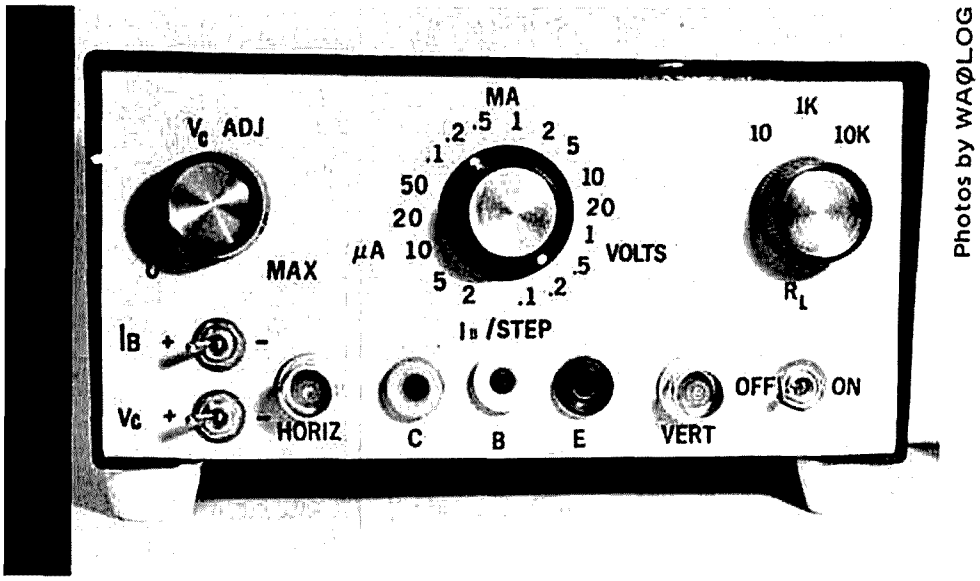
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ham radio



"Mother, I'm positive he's having an affair with another woman. I just heard him say Juliet. . .Sugar. . .Hotel. . .Whiskey. . . Two. . .Love."



Photos by WAØLOG

transistor curve tracer

This semiconductor curve tracer can be used with npn and pnp transistors, junction fets and mosfets, as well as diodes

Dan Wright, WA9LCX, 286 East 4575 South, Ogden, Utah 84403

Nothing is more frustrating than trying to design a circuit without the family of characteristic curves for the device you're going to use. In many instances the curves are not readily available or the device is unmarked. The simple-to-build, low-cost (less than $30) curve tracer shown in the photograph is the answer to this problem. It will display a family of collector characteristic curves for npn, pnp, jfets (P and N channel) and mosfets. It can also be used to display the volt/current characteristics of two-terminal semiconductors.

circuit

To produce the family of curves it is necessary to vary the base voltage in discrete steps while sweeping the collector voltage from zero to maximum at each step. As shown in fig. 1, the collector voltage is a 120-Hz rectified sine wave from a bridge rectifier, CR10-CR13. The maximum collector voltage is varied by R36 and the proper polarity is selected by S5.

The base voltage steps are synchronized to the 120-Hz collector voltage by Q1, Q2 and Q3. Transistor Q1 is an input amplifier that squares up the rectified sine wave. Q2 and Q3 form a one-shot

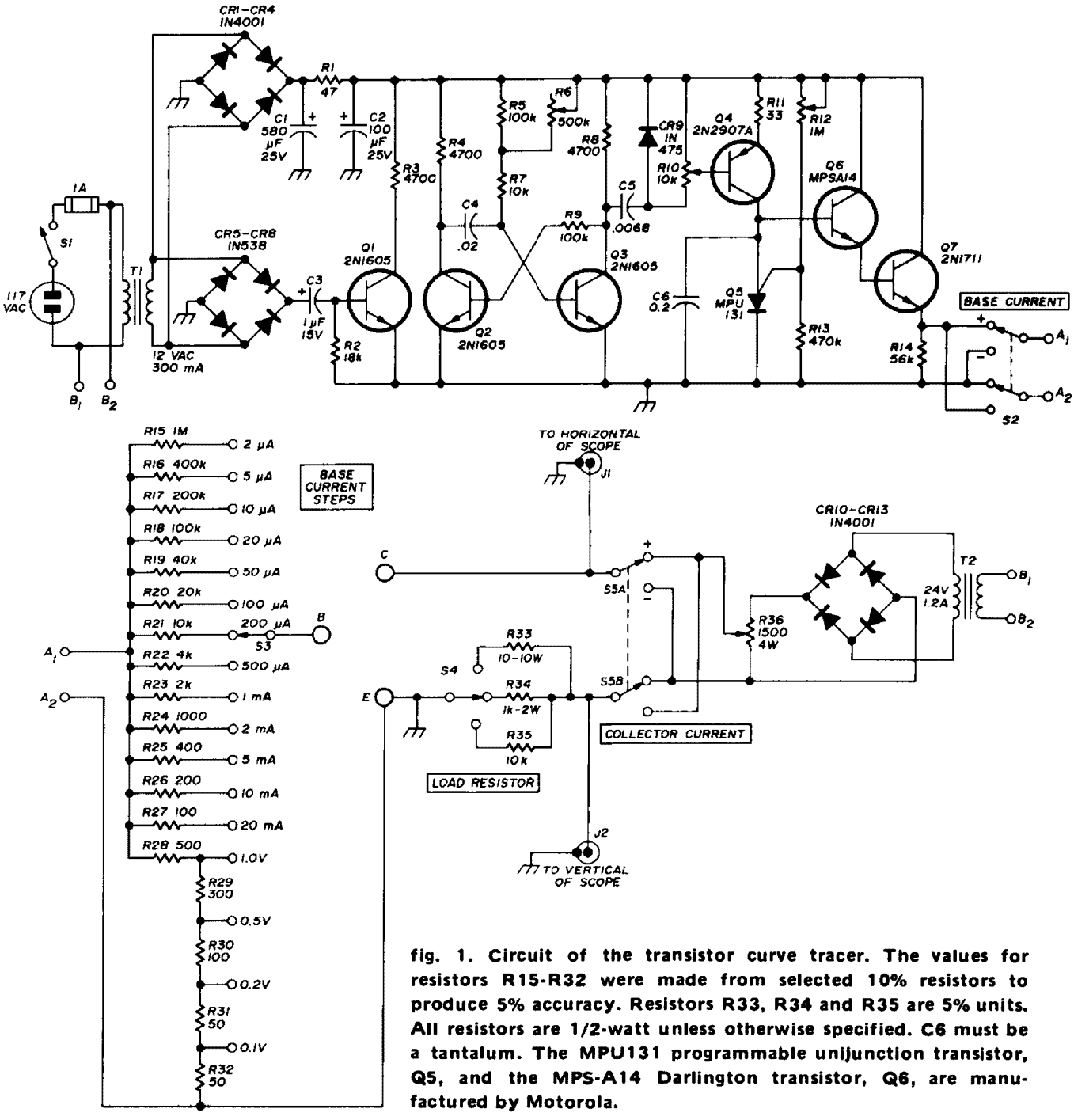


fig. 1. Circuit of the transistor curve tracer. The values for resistors R15-R32 were made from selected 10% resistors to produce 5% accuracy. Resistors R33, R34 and R35 are 5% units. All resistors are 1/2-watt unless otherwise specified. C6 must be a tantalum. The MPU131 programmable unijunction transistor, Q5, and the MPS-A14 Darlington transistor, Q6, are manufactured by Motorola.

multivibrator. The output pulse from the multivibrator is differentiated by R10 and C5. This pulse is synchronized with the beginning of the collector sweep by adjusting R6. Each time the step generator transistor Q4 is turned on by the synchronizing pulse, C6 receives equal current pulses, producing equal voltage steps.

The voltage between each step is controlled by R10. The programmable unijunction transistor, Q5, resets the stairstep generator back to zero. When the anode voltage goes higher than the gate voltage, the unijunction transistor fires and the generator is reset. The gate voltage is adjusted by R12. The stairstep voltage waveform is coupled to the selectable bias resistors (R15 - R32) for the test transistor by Q6 and Q7. The base bias is selected by S3 and the proper bias polarity is selected by S2.

The collector voltage is measured directly and applied to the horizontal

input of the oscilloscope. The collector current is determined by measuring the voltage across one of the load resistors (R33, R34 or R35) and applying it to the vertical input of the scope.

selecting and combining 10% resistors to obtain the necessary 5% tolerance.

Alignment is fairly simple. First, preset R12 to minimum resistance. Monitor the step voltage output at the emitter of Q7

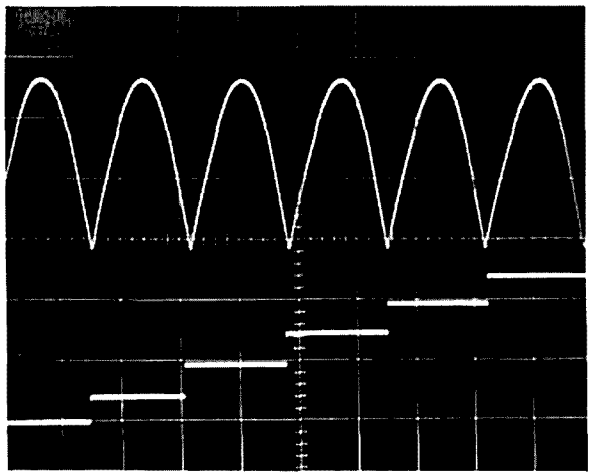


fig. 2. This dual-trace oscilloscope display shows the relationship between the base and collector voltage waveforms.

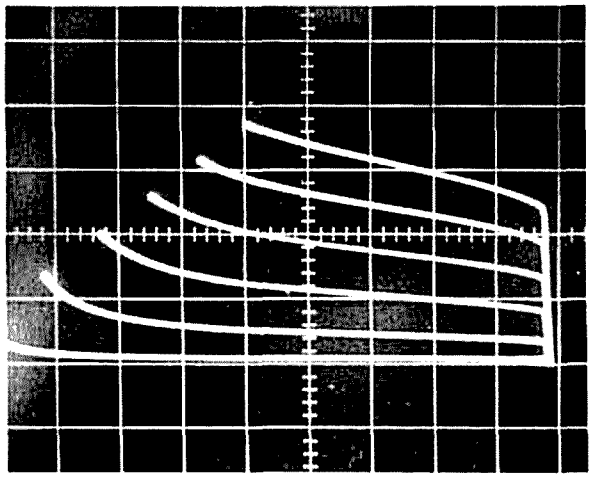


fig. 3. Curves for a 2N404 pnp transistor. Horizontal sensitivity is 2V/cm and vertical sensitivity is 2mA/cm. Base current is 5 μA per step.

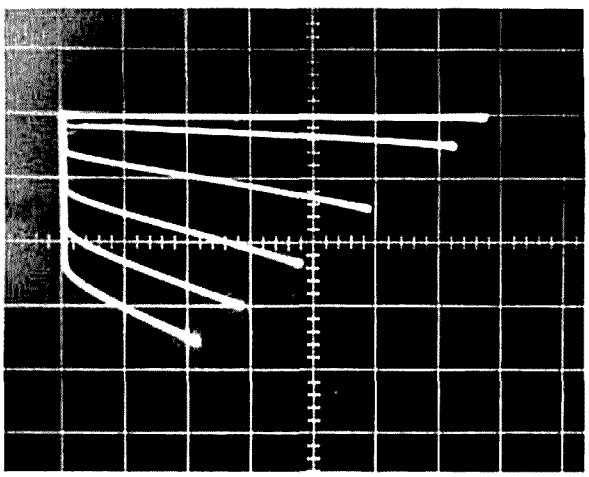


fig. 4. Curves for a 2N3055 npn power transistor. Horizontal sensitivity is 10V/cm and vertical sensitivity is 100mA/cm. Base current is 0.5 mA per step. The curves for all npn transistors will be inverted unless your oscilloscope has the capability to reverse signal polarity.

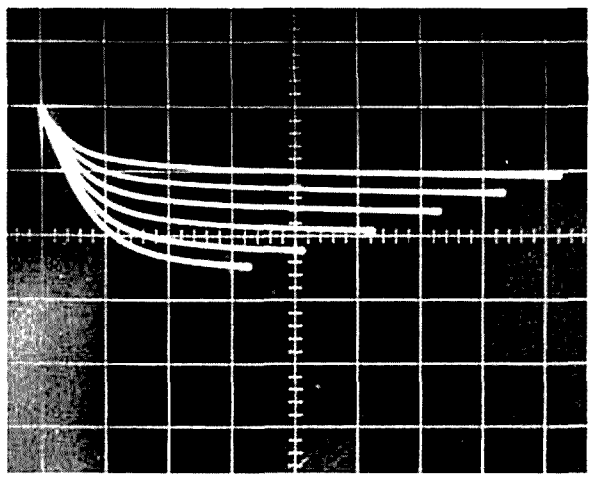


fig. 5. Characteristic curves for 2N4416 n-channel fet. Horizontal sensitivity is 2V/cm and vertical sensitivity is 5mA/cm. Gate voltage is +0.2 volts per step. Switch S2 can be used to provide characteristic fet curves with negative values of gate voltage.

construction and alignment

The curve tracer can be built on a perforated board or a printed circuit board, depending on the builder. Parts layout is not critical although it seems logical to keep all lead lengths short to minimize any stray signal pick-up. The base resistors (R15 - R32) are made by

with a scope and adjust R10 to produce the first step at approximately 2.2 volts. Adjust R12 to produce six steps before Q5 resets the generator. Adjust R6 so the base steps are synchronized with the collector waveform as shown in fig. 2. If a dual-trace scope is not available, this alignment can be made by adjusting R6 for the least amount of clutter on the

scope when a transistor is being tested on the curve tracer.

parts substitution

Before starting a project such as this, you always wonder if your junkbox parts can be substituted. This circuit seems to be very forgiving. Almost any npn transistor can be used for Q1, Q2 and Q3 (I used 2N1605s because that is what was in my junkbox). Almost any pnp transistor can be used for Q4. If you replace Q7, be sure to replace it with one of equal power dissipation and collector current.

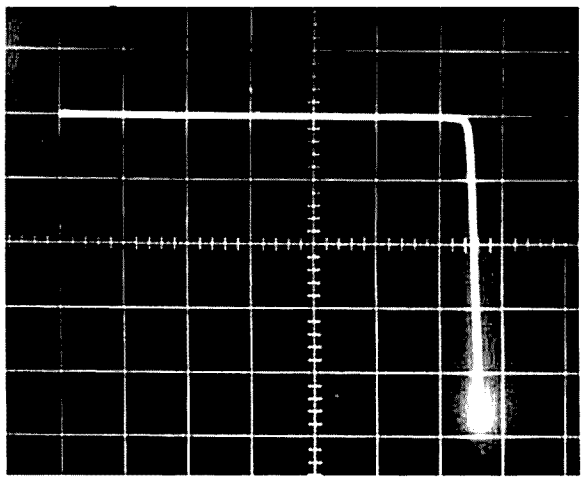


fig. 6. The curve tracer can also be used to display the voltage-current characteristics of a two-terminal semiconductor, such as a diode. This is the characteristic curve of a 1N429 zener diode. Vertical sensitivity is 2mA/cm and horizontal sensitivity is 1V/cm.

I do not recommend any substitutions for Q5 or Q6. They seem to work better than any of the other devices I tried, and they only cost 80 cents each from the surplus houses. The diodes can be substituted as needed so long as the replacements have similar PIV and forward current characteristics.

operation

Operation of the curve tracer is straightforward. First, select the proper base current and collector voltage for the device being tested. Select a load resistor, 1000 ohms for most small-signal devices, 10 ohms for power devices or 10k when you want to limit current for breakdown tests. Adjust your scope for a horizontal sensitivity of 2 volts/cm (this setting is convenient for most cases). The vertical sensitivity can be computed from the formula

$$I_c/cm = \frac{\text{sensitivity/cm}}{R_L}$$

Now, connect the transistor to the curve tracer. For fets the C terminal is connected to the drain, the B terminal is connected to the gate and the E terminal is connected to the source. Fig. 3 through fig. 6 show actual scope displays for various devices that were tested on this curve tracer.

conclusion

The completion of this project is only the beginning. If you consult a good textbook on transistors the meaning of these curves becomes much more explicit. If you do much work with semiconductors the information gained from these curves will save you countless hours of trial and error. I wish to thank my wife for her encouragement throughout this project.

ham radio



"Get that thing out of here!"

designing impedance-matching systems

A graphical method of designing impedance-matching networks for your favorite antenna

Bob Baird, W7CSD, 3740 Summers Lane, Klamath Falls, Oregon 97601

There's a line from Porgy and Bess, "the things that you're liable to read in the Bible — they ain't necessarily so." It's also a fact that the things you read in a good antenna theory book ain't necessarily so when it comes to your particular antenna.

What is the impedance of a vertical antenna? Well, you can look up the impedance on carefully constructed charts with electrical height in one direction and ohms in the other and get some ball-park estimates for both resistance and reactance. But what is the impedance of *your* vertical? What is its electrical height and is it the same for a 3-inch pipe as it is for a piece of wire? Down at the bottom of the page containing the graph you may find a footnote that says, "over a perfect ground." (Sometimes this footnote appears 3 pages removed from the chart.) Now, do you have a perfect saltwater ground or do you have a piece of wire ten-feet long or do you have the body of a motor vehicle?

There are just too many variables to come up with an answer either by eye balling or by taking physical measurements. To really know, you need to take some electrical measurements. And, to get any kind of precision, you need a good rf bridge. A $1000 rf bridge is not available to everybody but I had the use of one for a short time and came up with some rather surprising results.

mobile-mounted vertical

This article will involve itself with a couple of outstanding examples, but to start with, I was motivated by the results I obtained from a vehicle-mounted vertical. I don't care much about mobiling but I do like to ham from a camp location once I get there. If there are trees around, a good dipole is hard to beat. I get mine up by shooting a fish sinker, attached to a spinning outfit, with a slingshot. These surgical rubber tubing slingshots (available in any sports store) will shoot a one-ounce sinker over a fifty-foot-high limb quite easily. Then you reel the fish line back with a nylon cord attached, and raise your antenna.

But in some places, like the seashore, trees are not handy. So, I wanted a good vertical whip, a full quarter-wavelength high on 20, 15 and 10 meters. This was made from telescoping aluminum tubing.

The time-honored method of grid dipping to achieve resonance was followed to get the required length on the three bands.

I have a Ford van station wagon which has quite an expanse of roof for a ground plane. I looked in the good book for an approximate resistance for a quarter-wave ground plane and came up with the magic number of 36 ohms. "Aha, two hunks of 72-ohm coax in parallel should yield 36 ohms." Well, maybe it does, but my swr was higher than a cat's back and adjusting the telescoping whip to some point off resonance didn't help at all.

Finally, I took off one of the pieces of coax and things were just as bad as before (this was on 20 meters). In desperation I got a piece of 50-ohm coax and everything was hunky-dory. Then, I went to 15 meters — with 50-ohm line the swr was way too high. I changed back to 72-ohm coax and got on board. I had the same experience on 10 meters.

impedance measurement

Well, I finally got everything working all right, but I wondered just what was really going on, anyway. So, I borrowed an rf bridge and made some measurements. The General Radio 1606A rf bridge has two dials that read out resistance and reactance. At resonance the so-called j-factor or reactive component should read zero. With the vertical adjusted to the proper height in the middle of the 20-meter phone band I obtained a readout of 52 ± j0. On 15 meters the reading was 70 ± j0, and on 10 meters, 75 ± j0. All of these measurements were for one-quarter wavelength height. The differences in R seems to be attributable to the mounting (one corner of the van) and the extent of the ground plane (vehicle body). They would, of course, differ for every installation and every vehicle. However, I suspect that a lot of installations would work best with a 50-ohm line on one band and a 72-ohm line on another.

Measurements were actually made at the end of an electrically measured half wavelength of line in each case, in order to remove the equipment from proximity to the vehicle.

vertical antenna

In the W7CSD part of the world, crops are irrigated, and a 40-foot length of aluminum irrigation pipe is pretty easy to come by. ($14.30 for 3-inch tubing.) I set a one-quart soft-drink bottle in concrete and mounted a 40-foot section of irrigation pipe on top of it guyed with polyurethane rope (the kind used for water skiing). I thought this would work pretty well on 20 meters and also 40 and 75. The big problem was building matching networks for each band.



Author W7CSD with his three-band 40-foot irrigation-pipe vertical.

First of all, I needed to make some impedance measurements to see what I had. The first thing that I determined was that long radials will work fine on 20 meters, but short radials will not work well on 75. So, I used three 75-foot radials (actually, not really radial, but following convenient fence lines). More

would be better — the more the merrier. The same problems result as far as network design is concerned, whatever the ground system.

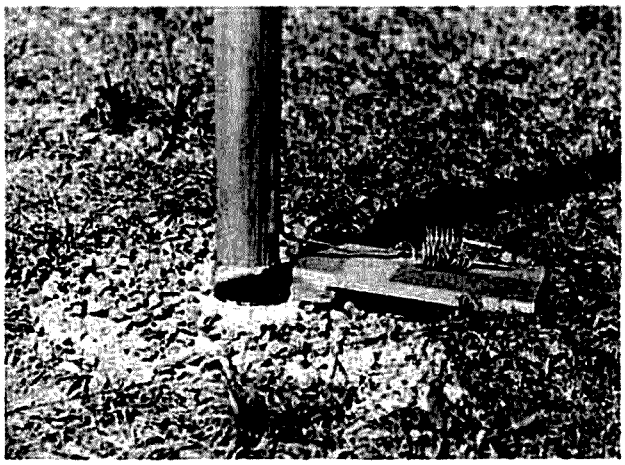
Forty feet on 20 meters is nearly 5/8 wavelength so you expect some kind of medium resistance value and capacitive reactance. 40 feet on 40 is more than 1/4 wavelength so you expect greater than quarter-wave resistance and inductive reactance. On 75 meters a 40-foot antenna is less than a quarter wavelength long so you would expect low resistance and a capacitive reactance. Actually, you need to know pretty close to the right values to be able to design the matching networks. With a ballpark estimate and a swr meter you might be able to get on board with some trial and error.

Taking advantage of the availability of the rf bridge again, I obtained the following measurements:

frequency	impedance
14.30 MHz	80 - j260 ohms
7.25 MHz	142 + j90 ohms
3.90 MHz	54 - j167 ohms

I wanted to match the above impedances to a 50-ohm line.

Matching a complex impedance to 50-ohm line with a T- or L-network by analytical methods is a lot of work. However, George Frese, A Consulting Radio Engineer in Wenatchee, Washington (ex W7FMI) came up with a fairly simple graphical solution to this type of problem.[1] Non-resonant vertical antennas are typical in a-m broadcasting.



Experimental loading system for using the 40-foot vertical on 75 meters.

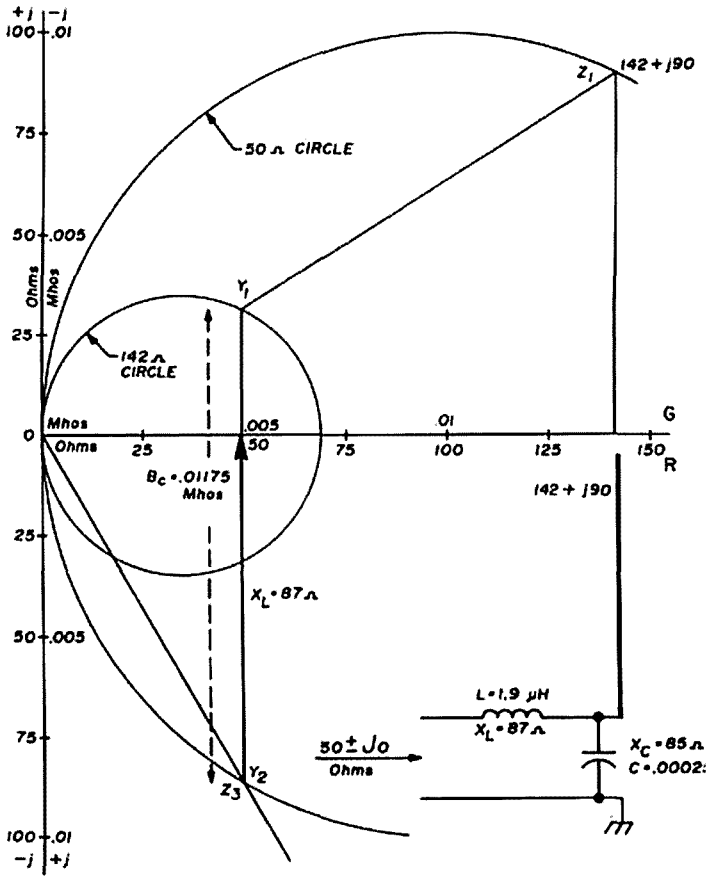


fig. 1. Graphical solution of an impedance-matching network for a 40-foot irrigation pipe vertical for operation on 7.25 MHz.

The graph has two calibrations, one for impedance in ohms and one for admittance in mhos. The two vertical calibrations have reversed signs; i.e., reactance going up is a +j, susceptance going up is -j (see fig. 1). Choice of scales is determined by the characteristics of the line and the antenna.

example 1

Let's take the 40-meter situation above as an example. I want to match 142 + j90 ohms to a 50 ± j0 transmission line. Since I have 142 ohms of resistance, the horizontal axis should go out to about 150 ohms. Likewise, the vertical should go to ± j100. The transmission line has an R of 50; 1/50 = 0.02 mhos = diameter of circle corresponding to 50 ohms. Therefore, 0.01 = radius of 50-ohm circle, needs to be on the paper, 20 squares = 0.01 seems appropriate, fig. 1.

1. With a good compass, draw the 50-ohm circle, radius = 0.01 mhos.

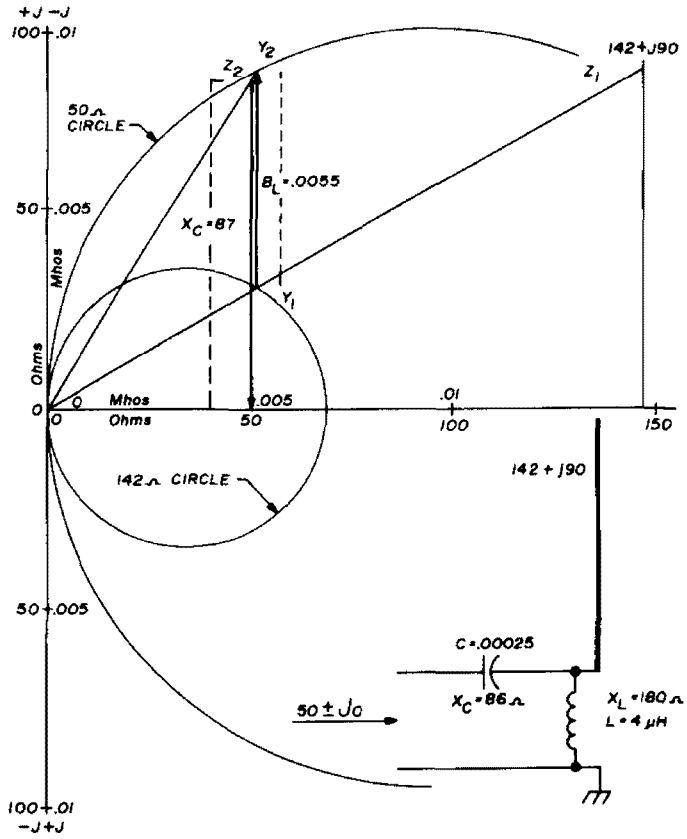


fig. 2. Another graphical solution for an impedance-matching network for a 40-foot irrigation pipe vertical for use on 7.25 MHz. Compare this network with the one shown in fig. 1.

2. Draw the 142-ohm circle, diameter = 1/142 = 0.00705, radius = 0.00352.

3. Point Z1 = 142 + j90 = 142 to right and 90° up.

4. Draw line from Point Z1 to origin.

5. Transfer Z1 to Y1 along this line to the 142-ohm circle.

6. Draw vertical line from point Y1 down to Y2 on the 50-ohm circle. The distance Y1 to Y2 is the susceptance of Y_c = +jB = 0.01175 mhos or X_c = 85 ohms. C = 0.000256 μF at 7.25 MHz (use a 0.00025 mica).

7. Draw line from Y2 through origin.

8. From the origin go out 50 ohms on the horizontal and draw a vertical line to the intersection of the line drawn in step 7. (In this case it just happens to coincide with line drawn in step 6.) This vertical distance (actually from Z3 to the horizontal) is X_L = 87 ohms; L = 1.9 microhenries.

Now, in case you are suspicious of all this hocus-pocus, let's take the result and work it out to see if it is true. Looking into the network, we are supposed to see 50 ± j0 ohms. Going out to the far end, we have the antenna, 142 + j90 ohms in parallel with the capacitor, 0 − j85 ohms. Using the formula for parallel impedances,

$$Z_{combination} = \frac{Z1\ Z2}{Z1 + Z2}$$

$$Z = \frac{(-j85)\ (142 + j90)}{-j85 + 142 + j90}$$

Changing to the polar form,

$$Z = \frac{(85 \angle -90^\circ)\ (168 \angle 32.4^\circ)}{142 \angle 2^\circ}$$

$$= 101 \angle -59.6^\circ = 51 - j87 \text{ ohms}$$

Combining the series inductance, X_L = 87 ohms, we obtain 51 − j87 + j87 = 51 ± j0 ohms. Accuracy is as good as the graphical method. Greater accuracy can be obtained from a larger graph.

The above solution is the one you want for two reasons. First of all, the network should look like a low-pass filter which will discriminate against harmonics. Also, you can eyeball the 75-meter situation and see that all you need is a series inductance, X_L = j167 ohms – the 54 ohms R is close enough. Likewise, the same kind of network is desirable for 20 meters so that the same coil, tapped at the proper places, can be used for all three bands. However, there is another solution which, for another problem, might be more desirable.

example 2

Steps 1 through 5 same as in example 1.

6. Draw vertical line from point Y1 *up* to Y2 to the 50-ohm circle. This vertical distance (in the up direction = −jB, is the susceptance Y_L = −jB = 0.0055 mhos or K_L = 180 ohms; L = 4 microhenries (see fig. 2).

7. Draw line from Y2 through the origin.

8. From the origin go out 50 ohms horizontal and draw a vertical line to intersect the line drawn in step 7. (This line happens to nearly coincide with the line of step 6.) This vertical distance equals 87 ohms and is downward going so has capacitive reactance; C = .00025 μF, approximately.

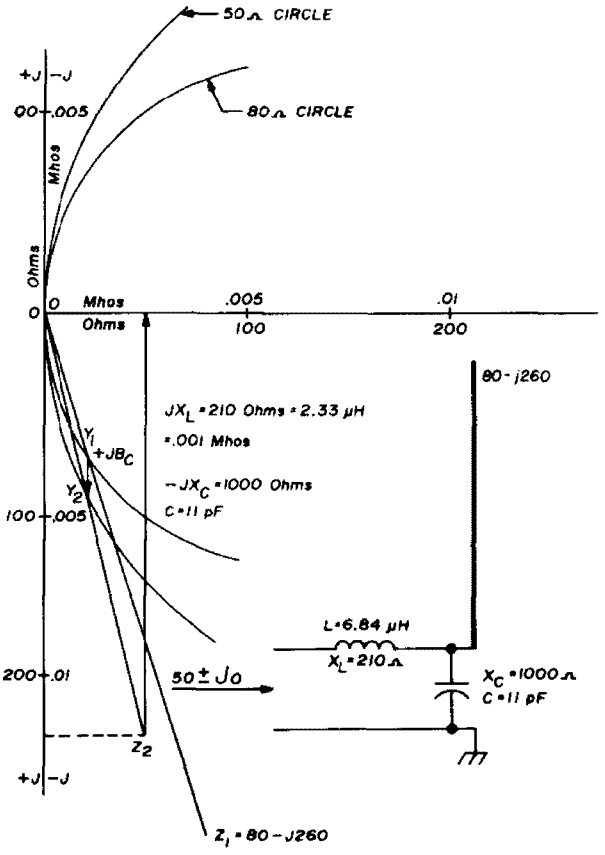


fig. 3. Graphical solution for an impedance matching network to the 40-foot vertical on 14.30 MHz. This network can be combined with the network in fig. 2 to provide the multiband system shown in fig. 4.

Working the problem again, the 142 + j90-ohm antenna is in parallel with the inductor X_L = j180 ohms

$$Z = \frac{(j180)\,(142 + j90)}{j180 + 142 + j90}$$

$$= \frac{(180 \angle 90^\circ)\,(168.5 \angle 32.4^\circ)}{305 \angle 62.3^\circ}$$

$$= 99.5 \angle 60^\circ = 49.8 + j86.2 \text{ ohms}$$

Combining with the series capacitor

Z_{in} = 49.8 + j86.2 - j87 = 49.8 - j0.8 ohms

Both answers should be 50 ± j0, but due to the inaccuracies of graphic construction, X_L in example 1 appears to be a little to the right of where it should be, and X_C in example 2 seems to be to the left of where it should be. Either way, you are within ±2%. It is doubtful that you will be able to get a capacitor or wind an inductor within this tolerance.

example 3

Now, let's look at 20 meters (14.3 MHz). You can use the same admittance scale but the impedance scale will have to be one square = 10 ohms to get everything on the graph (fig. 3). Since you want capacitance to ground and a series inductor, raise the horizontal axis and work below the line only. The distance Y1 to Y2 is very short and, hence, of questionable accuracy. If you had used the other solution and gone up instead of down, Y1 to Y2 would have been much greater and much more accurately measurable. But, for the reasons mentioned before you do not want this solution.

So, as closely as you can measure, Y1 to Y2 = 0.001 mhos or X_C = 1000 ohms and C = 11 pF. Locating Z2 is a little uncertain on this graph, too, but it comes out with X_L = 210 ohms and L = 2.33 microhenries. If you care to work out the problem again you will get Z = 48 + j8. This is still pretty close to 50 ohms.

proof of the pudding

Two months transpired between making the measurements and building the matching networks, and the commercial rf impedance bridge had been returned. With the aid of an L/C meter I wound and tapped a coil at 1.9, 2.33 and 6.84 microhenries. This could be done by any other method, including grid dipping using a known capacitor in a resonant circuit and solving for L. A 0.00025-μF mica capacitor was readily available. I used a small variable for the 11-pF capacitor.

On 75 and 40 meters I had an swr of

1.0:1 on the first trial. The swr on 20 meters was up near 1.5:1 and minimum was with the little 11-pF variable wide open. Removal of the variable capacitor and moving the inductor tap one-half turn yielded an swr of 1.0:1. Apparently, the capacitance of the coil to the aluminum box housing the network was very close to the required 11-pF. The final circuit of the network is shown in fig. 4.

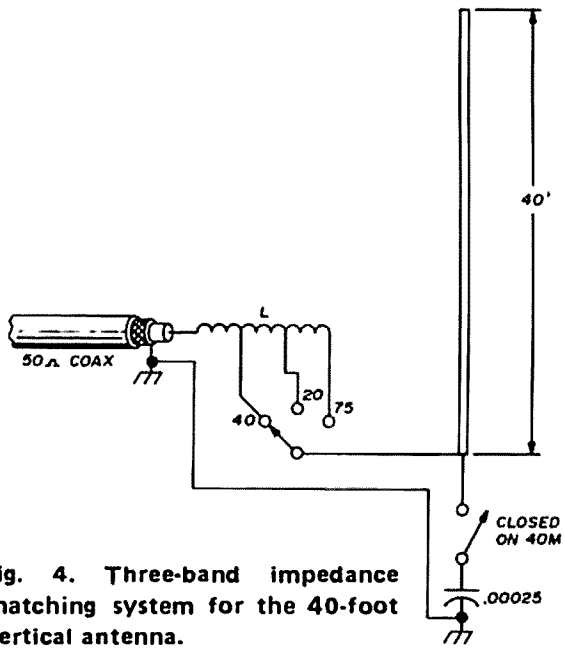


fig. 4. Three-band impedance matching system for the 40-foot vertical antenna.

summary

I have received excellent reports on all bands. I have no comparison on 75 and 40. On 20 the vertical cannot compete with a cubical quad, but this is not surprising. However, on DX contacts it is only down about 1 S-unit.

The big problem is how to measure the impedance of the antenna without a $1000 bridge. I think I may have some answers for this which I will verify when antenna weather comes again. There are several possibilities that are close enough that little trial and error would be necessary to get everything tuned up.

reference

1. George Frese, "Graphical Solutions to RF Networks," *Broadcast Engineering,* January, 1964, page 12.

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how to compare the efficiency of linear power amplifiers

Accurately comparing the efficiency of two linear amplifiers is not difficult, but it requires some care

Carl Drumeller, W5JJ, 5824 N.W. 58 Street, Warr Acres, Oklahoma 73122

While listening on an amateur band, I overheard one amateur telling another his results in comparing the efficiency of two commercially-built rf power amplifiers. Each of these amplifiers used vacuum tubes, and was designed for amplifying the output of ssb exciters. The comparison was based solely upon the ratio of dc power input to rf power output while speaking the same words into the exciter's microphone at approximately the same amplitude.

This seems to be a commonly-used and more or less universally accepted means of evaluating the efficiency and effectivity of rf power amplifiers used as linear amplifiers. But is it a valid comparison?

In the particular instance in question, one rf power amplifier showed an efficiency of 67%, the other 40%. The high apparent efficiency of one is an immediate *flag* for suspecting the validity of the test. That figure, 67%, borders upon the theoretical maximum that a vacuum-tube amplifying stage can produce in linear service. It supposes that every parameter is at its perfect peak of optimum adjustment, that associated circuits are without loss, that nirvana has been achieved. The other figure, 40%, is a much more believable one!

the comparison

What, then, constitutes a valid means of comparing the efficiency of two linear amplifiers? The clue lies in that term, *linear.* Vacuum-tube amplifiers can be designed and constructed to be amazingly efficient as converters of dc power input into rf power output. By running dc grid bias very, very high (many times the cutoff point) and running the rf grid excitation high enough to saturate the tube, an efficiency of around 90% can be achieved.

But, would you want it? You probably wouldn't, for that rf output would contain as much power in harmonics as in the

fundamental. This is not precisely what you would desire if you wanted to keep on good terms with the FCC. Nor is it what you would want for amplifying any form of amplitude-modulated signal, and, of course, ssb is one form of amplitude modulation.

Leaving such an extreme, really a class-C stage, and looking toward class B or AB, the classes usually associated with linear amplification, you do not find a high degree of efficiency. The true efficiency, the conversion of dc input power into rf output power of the desired frequency, is not significantly less than that achieved with a class-C stage. What you do find, however, is the possibility of approaching truly linear amplification, amplification in which the output waveform is a true reproduction (in every manner except amplitude) of the input waveform. And this is what you desire when you build, buy, or tune up an amplifying stage connected to the output of a ssb exciter. A serendipitous side effect of such a stage is a satisfying reduction in the generation of harmonics.

how to do it

This suggests some guidelines for comparative tests. You would like to set some limits on just how much of a departure from linearity you will tolerate in the interest of efficiency. Manufacturers do this sort of thing; they come up with a figure like, say, -35 dB for third-order intermodulation products.

Like most trustworthy and informative measurements, such a measurement requires rather expensive tools. In this case, the tools are a two-tone oscillator (really neither difficult to build nor expensive to buy) and a simple spectrum analyzer, such as the Heathkit SB-620. The two-tone audio-frequency signal generating device needs to embody a low duty cycle pulser, something that will let the signal through for a third or less of the time. This permits running the amplifier under test at full load for moderate periods of time without cooking the tubes.

Of course, for any power output test you have to have a dummy load/rf wattmeter. And for the dc input measurements you have to have the needed voltage and current meters.

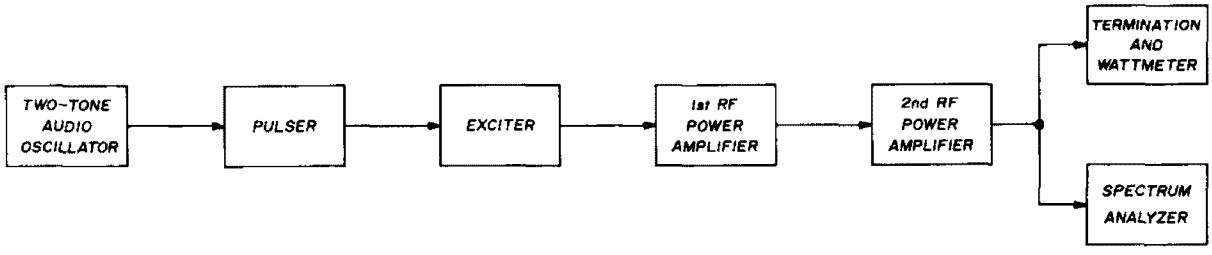


fig. 1. Equipment arrangement for comparing the efficiency of two rf power amplifiers. The amplifiers are switched in and out of the line with their own built-in bypass relays.

As most linear amplifiers are built with integral dpdt bypass relays, a convenient way of making a comparison test is to arrange the two linear amplifiers in series, following the exciter. The control actuating the in-circuit or out-of-circuit status of the bypass relay can then be used to instantly select the particular amplifier to be observed. Fig. 1 shows a block diagram of the necessary arrangement.

linearity

Since it's highly probable that any amplifier approaching true linearity will have low harmonic output, it is not necessary to use a wideband spectrum analyzer to check the relative (or the absolute) power contained in harmonics. Only a narrow-band analyzer is needed. The odd-order, or intermodulation, distortion products shown in fig. 2 can be observed only with an analyzer that has fairly good resolution. However, such resolution is not absolutely necessary; an ideal of relative magnitude of, say, $2f_2$-f_1

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Triacs — thermo tab package — 1 amp. - 400 v., $.80 ea.; 3 amp. - 400 v., $1.40 ea.; 4 amp. - 200 v., $1.20 ea.; 6 amp. - 200 v., $1.40 ea.; 6 amp. - 400 v., $1.60 ea.; 6 amp. - 500 v., $1.80 ea.; 8 amp. - 200 v., $1.60 ea.; 8 amp. - 500 v., $1.80 ea.
To-5 case, 1 amp. - 200 v., $.70 ea.; 1 amp. - 400 v., $1.00 ea.
SCR 200 v. - 8 amp. thermo tab ..... $.80 ea.

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as compared with either f_1 or f_2 (which probably will be seen as a single blip) can be quite definitive. By making a note of this relationship and then shifting over to the other amplifier, you can form an accurate idea of their relative linearity.

Having established an idea of relative linearity, the next step is to optimize the linearity of each amplifier by careful adjustment of tuning, loading and rf

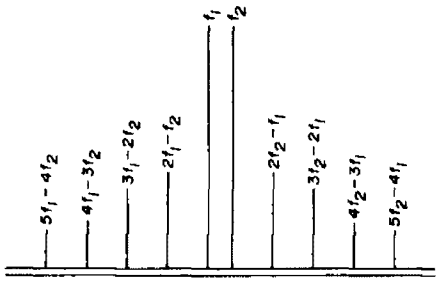


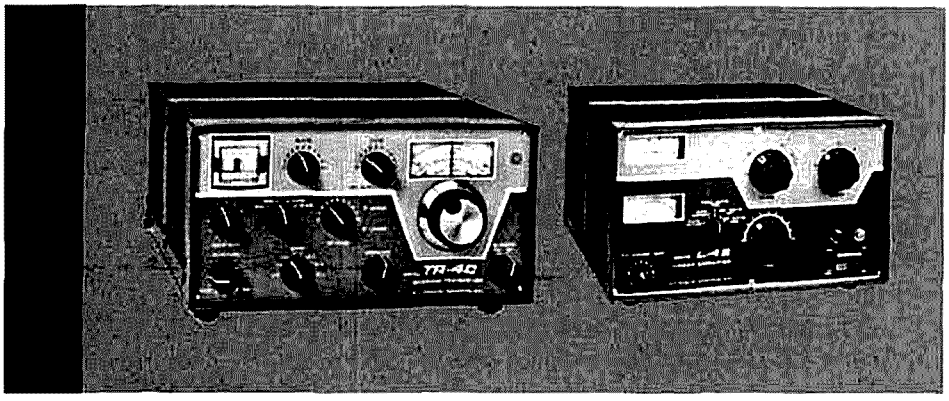
fig. 2. Spectrum of odd-order distortion products in an rf amplifier with input frequencies f1 and f2.

excitation. Then, make both amplifiers exhibit equal linearity by decreasing the rf excitation to the amplifier having the greater distortion. Measure its dc power input and rf power output. Go to the other amplifier, remembering to restore the rf excitation to its optimum value, and measure its input and output. Then and only then are you in a position to talk about the relative efficiency of two different rf power amplifiers.

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ham radio



ham radio sweepstakes winners

WA9YII wins grand prize, puts new Drake TR-4C hf transceiver and L-4B linear on the air; WB4KIT is lucky winner of Robot sstv camera and monitor

Skip Tenney, W1NLB, Publisher, Ham Radio Magazine

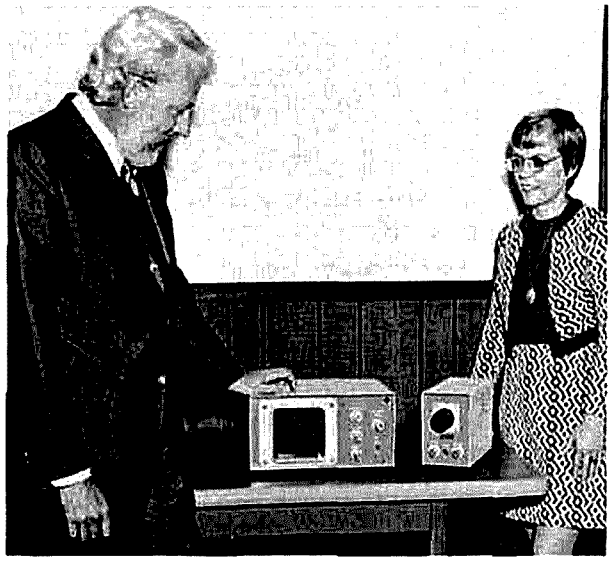
It was more work than ever, opening ten's of thousands of your letters and trying to keep our records straight. However, the *1973 Ham Radio Sweepstakes* is now history, and our local post office can take a breather and so can we.

This year's contest seems to have created more interest than ever before, not only through the mail, but also at many of the Hamfests we have visited this year such as SAROC, the Tropical Hamboree, the Dayton Hamvention and a number of others.

Our Grand prize winner was Randy E. Thompson, WA9YII. He won a brand new R.L. Drake TR-4C Transceiver and AC power Supply plus a Drake L4-B Linear Amplifier. Randy tells us that although he is now a Technician he is going for his Advanced license in the very near future. He is in the graduating class at the Vocational-Technical Institute of Southern Illinois University and has accepted a position with the National Accelerator Laboratory near Batavia, Illinois.

The new TR-4C offers everything that made it's predecessor, the TR-4, so popu-

Jim Fisk and Pat Hawes of the Ham Radio editorial staff look over the Robot SSTV Camera and Monitor won by WB4KIT.



lar, plus a number of new features including 1-kHz dial calibration. This versatile transceiver provides ssb, CW and a-m operation on all amateur bands from 80 to 15 meters plus 28.5 to 29.1 MHz. Accessory crystals are available to cover the complete 10 meter band.

Of course, the L4-B is one of the best known 2-kW amplifiers in the business. Rugged and dependable, it features a trouble-free, conservative design. Randy should be set for many years of service from this fine equipment. The L4-B uses a pair of Eimac 3-500Zs in a class-B grounded-grid circuit featuring a broadband-tuned input, negative rf feedback and transmitting agc for higher audio level without clipping.

The second prize, a Robot Slow-Scan Television Camera and Monitor, created much interest from many of our entrants. This is the same package that has added a whole new dimension, that of sight, to amateur radio for so many operators all over the World. The lucky winner of the Robot Model 70 SSTV Monitor and Model 80 SSTV Camera was John S. Harvey, WB4KIT. We'll all be "looking" for him on the air very soon.

Both the Robot Camera and the Robot Monitor feature an all solid-state design except for the picture and camera tubes. Particular design attention has been given to making these units easily adaptable to virtually any ssb station. A couple of patch cables, one to the microphone jack and one to the audio output, is all that is necessary to connect the equipment to your rig, and put it on the air.



Skip Tenney calls WA9YII to let him know he has won the Grand Prize — a new Drake TR-4C transceiver and L-4B linear. Randy, obviously very pleased, said, "This is the first time I've ever won anything!"

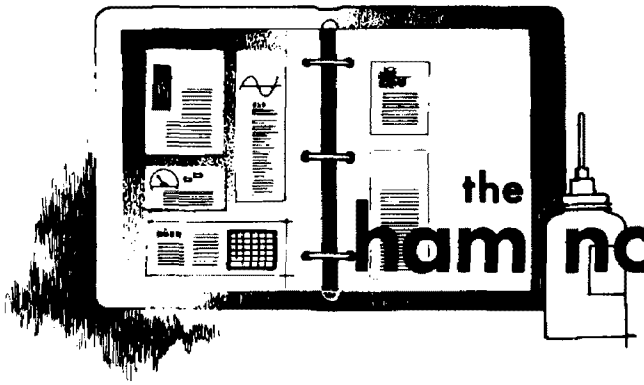
Skip Tenney, W1NLB, publisher of Ham Radio, picks the Grand Prize winner in the 1973 Ham radio Sweepstakes. Looking on are assistant publisher Hilda Wetherbee, editorial assistant Pat Hawes, WN1QJN, and editor Jim Fisk, W1DTY.



Versatility plus was the third prize. Each of our three third-prize winners received a Drake TR-22 Two-Meter FM Transceiver. This is that wonderful little unit that is equally at home over your shoulder, in your car or at home. It can operate on its own battery pack or on 12 Vdc or 115 Vac. John R. Low, K3YHR, Vinton A. Buffenarger, W6PSC, and Robert K. Jackson, WBØISI will be coming through on their local repeaters in the near future, and we're sure that they'll be saying good things about their nifty new rigs.

We're sorry that everyone couldn't win this year, but there are some exciting, new prizes that we're lining up for 1974. We hope that you'll plan on winning one of them and we will try our best to help you out.

ham radio



the ham notebook

arc suppression networks

Unless special precautions are taken, operation of a relay or switch in close proximity to sensitive electronic circuits is a potential source of trouble. This trouble generally results from arcs caused by opening of the relay contacts, causing transients that can easily interfere with the proper operation of sensitive circuits.

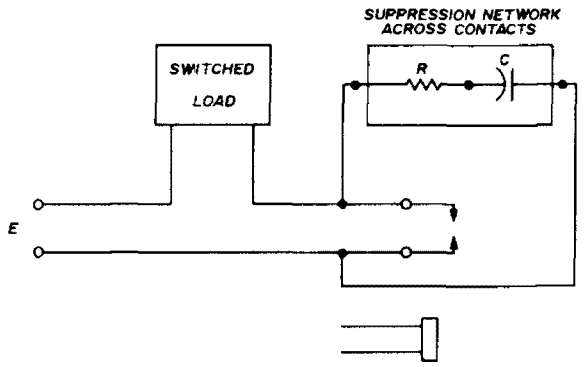


fig. 1. Simple arc-suppression network eliminates transients which can cause problems in sensitive circuits. Values of R and C may be determined with nomograph in fig. 2.

The effects of these arcs can be minimized by the installation of an arc suppression network directly across the relay contacts to absorb the energy which would otherwise be dissipated in the arc itself.

The network itself is simply a resistor and capacitor in series. The capacitor just "absorbs" the arc by charging when the contacts open (see fig. 1). The resistor limits the current generated by the discharging capacitor upon contact closure. Were it not for the resistor, severe contact pitting could result from large momentary discharge currents.

The optimum values of R and C can be chosen by the equation

$$R = \frac{E}{10\,(3.16\sqrt{C})^{\,1\,+\,50/E}}$$

where R is the resistance in ohms, C the capacitance in microfarads and E the open-circuit potential in volts. Obviously, this equation is rather unwieldy, and selection of values from it could prove to be quite a chore! For convenience, this equation has been solved graphically in fig. 2. By using this nomograph any number of R-C combinations can be chosen with ease.

Suppose, for example, that a suppres-

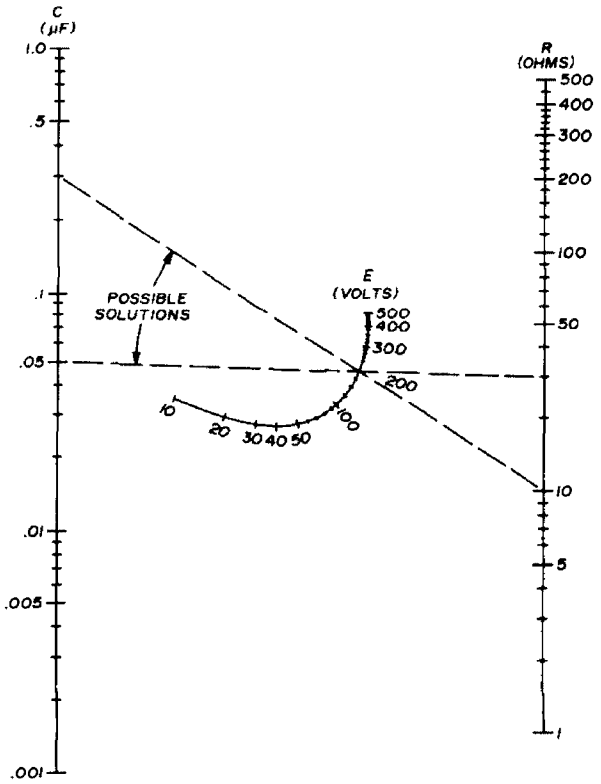


fig. 2. Nomograph for finding R and C values for the arc-suppression network shown in fig. 1.

sion network must be selected for a switched potential of 200 volts and that a 0.05-μF capacitor is readily available. Drawing a line from 0.05 on the C scale through 200 on the E scale yields a value of 30 ohms on the R scale. The network, therefore, will be the series combination of 30 ohms and 0.05 μF. An infinite number of other combinations of R and C are possible, and the final choice of components will probably depend upon what you have on hand. Another suitable choice here would be 0.3 μF and 10 ohms. This can be verified from fig. 2.

Since the nomograph is so easy to use, several combinations of R and C should be tried to find the one which is physically smallest and can be constructed with only two standard values. The result will be a network which will go a long way in eliminating headaches caused by contact arcing.

James McAlister, WA5EKA

pilot-lamp life

For some time, I have been experiencing relatively short life for the 120-volt pilot lights I use. Investigation showed that the green light had short life, but the high-voltage indicator (red bulb) seemed to last indefinitely. The reason? The red indicator came on by steps — that is, half-power, then full power.

An ohmmeter check showed that the hot resistance of the S6 bulb was 2400 ohms, while the cold resistance was only 200 ohms. This meant that the small bulb had to sustain a temporary power surge of 72 watts. I recalled that BC stations, where dependability is of great importance, have, for years, used a dropping resistance in series with most indicator lamps. The resulting setup at W2OLU calls for a dropping resistor of approximately 450 ohms. This is not at all critical — any value from 400 to 500 ohms will do. The resultant *hot* current is about 42 mA and the maximum surge is reduced to a small fraction of its former value.

Neil Johnson, W2OLU

pogo stick for reflex klystrons

Stabilizing klystrons is an important part of the microwave station. There are several ways of accomplishing this, but the following, I feel, is more advantageous than others. The primary problem is connecting the output of the afc network which is operating at low voltage levels to the high voltage levels of the klystron.

The output from the receiver discriminator is around zero. It is therefore logical to make the afc compatible. Type 741 op amps connected as an integrator, impedance matcher and inverter perform this function at minimum cost and parts.

Now the problem comes to light. How do we get this to the reflector of the klystron and do it inexpensively with semi-conductors? After all, excluding the klystron, the entire station consists of semiconductors. Let's not spoil it.

ICU1 is not necessary to the operation of this circuit (see fig. 3). However, with the afc disconnected, the voltage at point C is not controllable. The fact that the input to U1 is a voltage node and the loop is operating around zero means we can open the loop at point X and not appreciably change the output frequency of the klystron. The slight change that does occur is due to the input offset current of the 741 IC.

Let's disconnect R7 at point X and look at the operation of this circuit. The output of U1 is zero with R7 open or grounded. If point A is zero, and the base of Q1 is at -12 V, we must have approximately 1 mA through R1. If the emitter current of Q1 is 1 mA then the collector current is 1 mA; 1 mA through R2 gives us 75 volts across R2.

Note the voltage divider R3 and R4 which makes the base of Q2 -150 V. Therefore, the emitter must be the same, plus one diode drop, which is insignificant. With the emitter of Q2 at -150 V, 75 V across R2 and 12 V across R1, we are left with approximately 63 V across Q1.

The reason for Q2 is now evident. It is not necessary except that if only Q1 was used it would have to be a high-voltage device which implies high cost. As the circuit stands, Q1 and Q2 can be 100 volt devices and still maintain a good safety margin.

We established earlier that the current through R1 was 1 mA. Q1 and Q2 are in series with R1, thus the current in the series string is 1 mA. This results in Q2 and R5 having the same voltage drops as Q1 and R2 respectively. The voltage at point Y is now 300 volts. This leaves 300 volts across R6 and R9. R6 is a ten-turn Helipot for adjusting the voltage on the klystron repeller so it is operating in its proper mode (160 volt mode is used at this station).

Understanding that Q1, Q2 and the resistors constitute a series circuit is important. With this in mind let's put a positive voltage at point X. Point A goes negative (U1 inverts). With point A less than zero the current in R1 reduces, thus reducing the current in the series string. With a reduced current the voltage across R9 reduces, making point C more negative. With a negative voltage at point X, A goes positive. More current results in the string causing point C to become less negative (positive direction).

By changing R7, the gain of U1 changes, thus changing the loop transfer function (ratio of point-X voltage change to point-C voltage change). The value of R7 will thus be determined primarily by the transfer function of the particular afc circuit used. (In my station U1 is operated with a gain of −10.)

Resistor R1 is chosen so the voltage at point-Y is approximately −350 V (slightly more than 1 mA). This is to insure a good voltage swing at point C without causing point C to go above −300 volts. CR1 is for protection of the klystron. The repeller should never go positive with respect to the cathode.

This circuit is in use at two stations operating at 3335 and 3365 MHz with great results. The klystron is able to track over its entire electrical tuning range. Total cost was less than $5.

Francis E. Adams, W6BPK

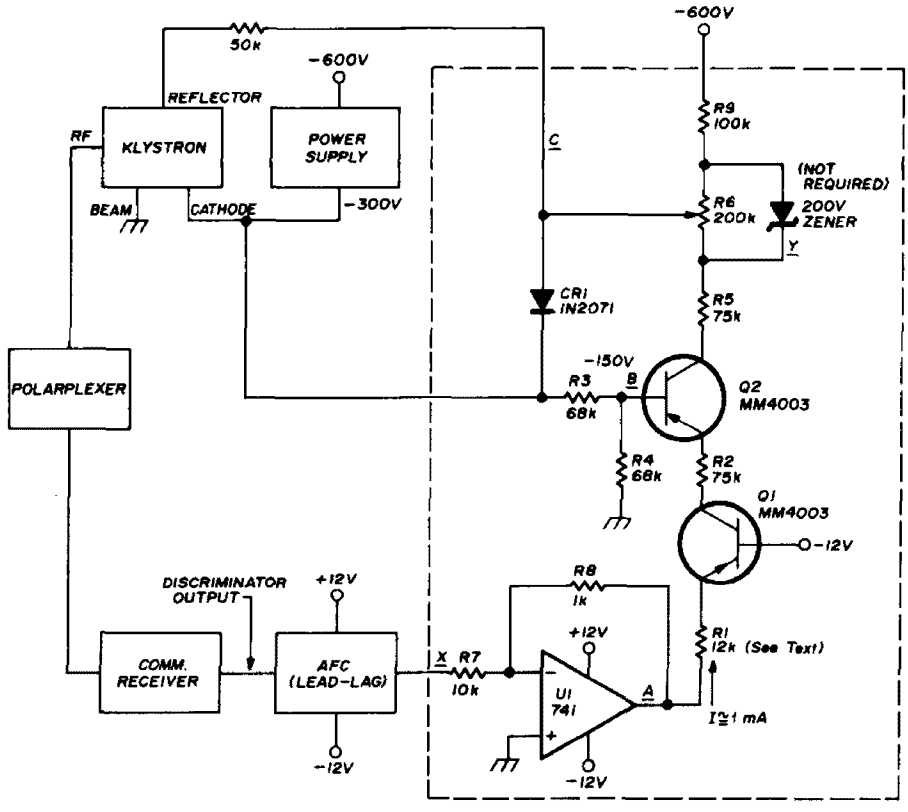
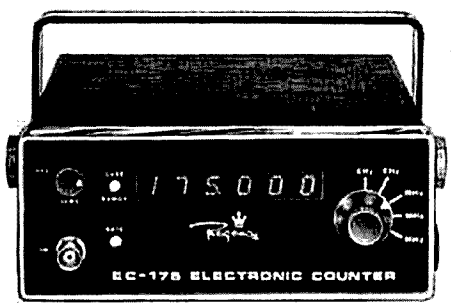


fig. 3. Solid-state voltage-tracking circuit for reflex klystrons.

new products

electronic frequency counter



Regency Electronics has moved in to a new area of electronic equipment with production of a six-digit electronic frequency counter. The all American made unit, the EC-175 Electronic Counter, is priced at $449.00

The unit's low price, combined with high stability, ease of operation and portability make the Regency Counter very desirable. The EC-175 is designed to enable the operator to measure crystal frequencies without mathematical computation. The counter reads out frequencies ranging from 5 Hz to 175 MHz. A five-position range switch with gate times of 1 ms, 10 ms, 100 ms, 1 second and 10 seconds allows direct measurement of any in-range frequencies to within 0.1 Hz.

The six-digit LED display features automatic blanking, automatic decimal point positioning and leading zero suppression. There is an overrange LED indicator for readings over 6 digits and separate LED indicator for count rate.

EC-175's proportional oven-controlled time base gives a short-term stability of 2 parts in 10^8 per day for FCC certification work. Long term stability is one part in 10^6 per 6 months. Temperature stability is 3 parts in 10^9 per degree Centigrade from zero to 50° C.

The Regency frequency counter has a built-in 100-kHz harmonic generator for direct calibration to WWV. A 10,7-MHz crystal oscillator for afc locking and i-f alignment work is also a built-in feature. A built-in mosfet preamp gives sensitivity of 100 mV at 100 MHz.

The Regency EC-175 Electronic Counter is now available from Regency distributors throughout the country. For more information, write to Regency Electronics, Inc., 7900 Pendleton Pike, Indianapolis, Indiana 46226 or use *check-off* on page 110.

keyer memory



A large capacity programmable-reprogrammable CW message memory, designed as a plug-in accessory for the EK-420 keyer, has been announced by Curtis Electro Devices. Aimed to satisfy the most demanding contester or traffic handler, the standard KM-420 offers a solid-state memory capacity of 1,024-bits — equivalent to 100 Morse characters. A second, options, 1,024-bit plug-in memory doubles this capacity. For maximum flexibility, the memory organization is switch selected to yield four different program arrangements.

Programs are selected and started by

More Details? CHECK–OFF Page 110

pushbuttons; terminated by a message pause or by manual break-in. A "reset" button instantly stops and resets the memory, a "hold" function allows manual insertion into a sequency. A 1 to 20 second repeat feature allows send and wait operation.

Memory programming is accomplished simply by sending the desired sequence in the "record" mode. Messages may be written as often as desired and stored indefinitely. Automatic and manual sending are indistinguishable. The KM-420 is priced at $299.95. The optional extra memory is $34.95. A remote control head for the Brown Brothers CTL Key is also $34.95.

For additional information, contact Curtis Electro Devices, Inc., Box 4090, Mountain View, California 94040, or use *check-off* on page 110.

vintage radio

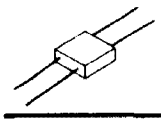
This new pictorial album of old radio equipment, edited by Morgan E. McMahon, should hold particular interest to amateurs and antique radio buffs who would like to know more about radio during the early part of the century. The book is full of interesting photographs and text describing much of the early apparatus used by such wireless pioneers as Marconi, DeForest and Sir Oliver Lodge. There's a complete section on early receivers including models manufactured by Grebe, Remler, Atwater-Kent and RCA; early vacuum tubes, coherers, crystal sets, headphones, horn speakers, test equipment and many special components. The list of 3000 radio models by manufacturer and year should be a particular aid to collectors who are having trouble dating a particular radio in their collection.

Soft-bound, 263 pages, hundreds of photographs and drawings, $4.95. Also available with hard cover for $6.95. Order from Vintage Radio, Box 2045, Palos Verdes Peninsula, California 90274. For more information, use *check-off* on page 110.

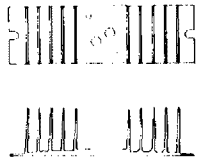
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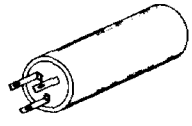
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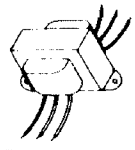
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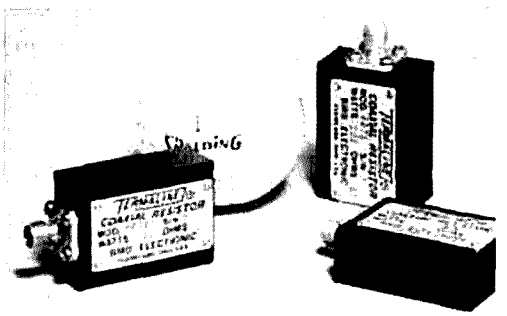
MECL data book

Motorola's new *MECL Data Book* is a fact book and a design guide wrapped in one cover. Basic performance specifications are provided for each IC device in the MECL II, MECL III and MECL 10,000 logic families. In 410 pages, complete dc, ac and performance data are given for each MECL product announced up to press-time. Convenient abbreviated guides offer a quick selector for MECL logic blocks.

In the 36-page general information section is contained a synoptic discussion of topics of value to the designer planning a MECL system: product performance data, design rules, system considerations,etc. In addition, dimensioned drawings of MECL IC packages are shown. References to the extensive MECL software support available are included by way of Motorola Application note abstracts, as well as references to a sampling of MECL articles in the trade press.

Motorola's *MECL Data Book* may be purchased for $2.00 per copy. Copies may be obtained by sending check or money order payable to Motorola, Inc. at Post Office Box 20924, Phoenix, Arizona 85036.

rf load resistor



A nearly hundred-to-one breakthrough in size reduction for a coaxial rf load resistor has been accomplished by Bird Electronic Corporation. The New *Termaline Miniload®* Model 8071 100-watt load weighs just 1¼ ounces and requires less than 3/4 cubic inch of space. Designed for reflection-free termination of 50-ohm coaxial systems, these dry miniature

More Details? CHECK–OFF Page 110

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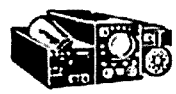
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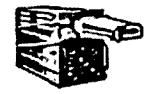
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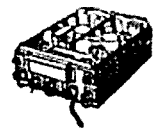


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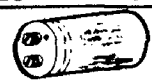
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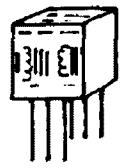


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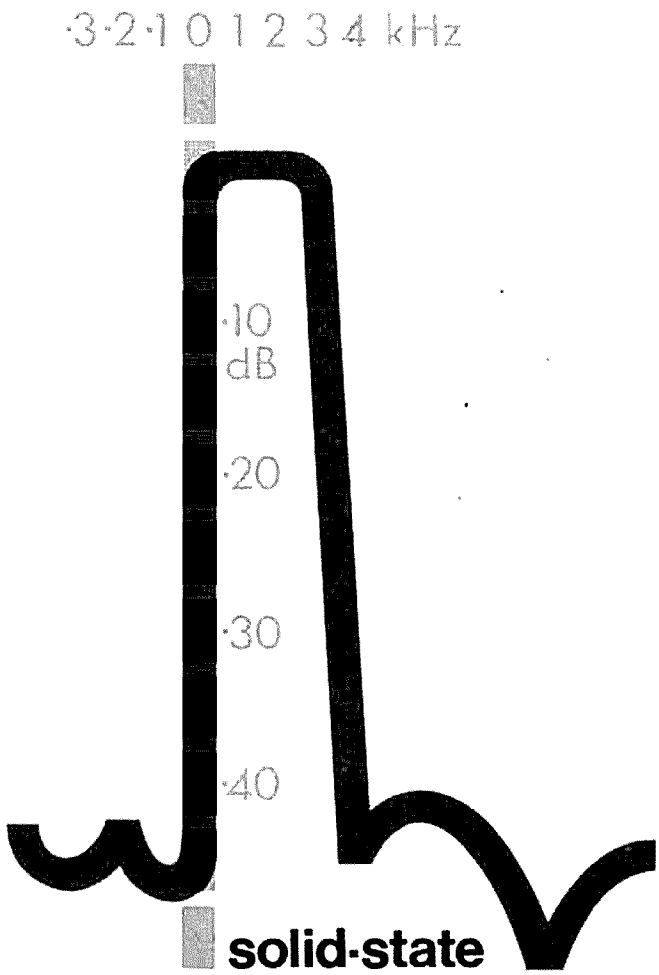
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ham radio magazine

AUGUST 1973



solid-state phasing-type ssb COMMUNICATIONS RECEIVER

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- frequency synthesizer 50

August, 1973
volume 6, number 8

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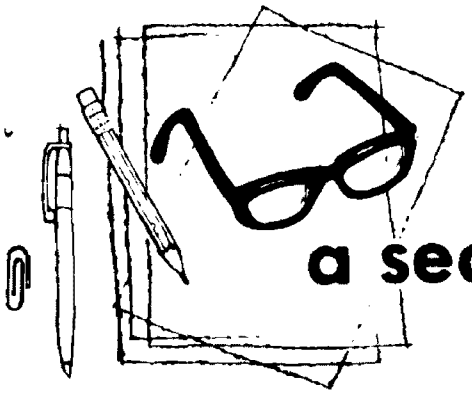
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a second look

by jim fisk

From all indications, it looks like there's a good possibility that we're going to lose the top 1 MHz of the amateur 220-MHz band to make way for a new Citizens Radio Service. Although this new CB service was originally proposed over two years ago by the Electronic Industries Association, the wheels of government grind ever so slowly, and it wasn't until early June that the FCC released their Docket 19759, proposing the creation of a new 224-225 MHz Class-E Citizens Radio Service. Comments are invited from interested parties (the usual original and 14 copies) before September 20, 1973.

According to a spokesman at the FCC, it is estimated that the new CB service could be finally approved by the Commission within 6 months to a year. The Commission also said that the new 1-MHz Citizens Band would be divided into 40 25-kHz fm channels. Eligibility would be similar to that for the present Class-D service on 27 MHz – any person 18 years or older who meets the basic criteria for licensing.

In proposing the new band, the FCC cited the need for additional frequencies to meet the requirements of the general public for improved radio communications and to relieve some of the heavy concentration of CB stations on the channels available in the 27-MHz band – the 49,000 Class-D licensees of 1959 had grown to nearly 900,000 in 1972.

Obviously, a new radio service such as this creates a huge new market for the electronics industry, and the EIA is pushing hard for early approval. Once the necessary equipment is mass produced, it is expected to sell for about $200 to $300 per unit; the EIA estimates that two to three years after the new service is started, it will represent a new electronics market amounting to some $300 million annually. That's a lot of boodle!

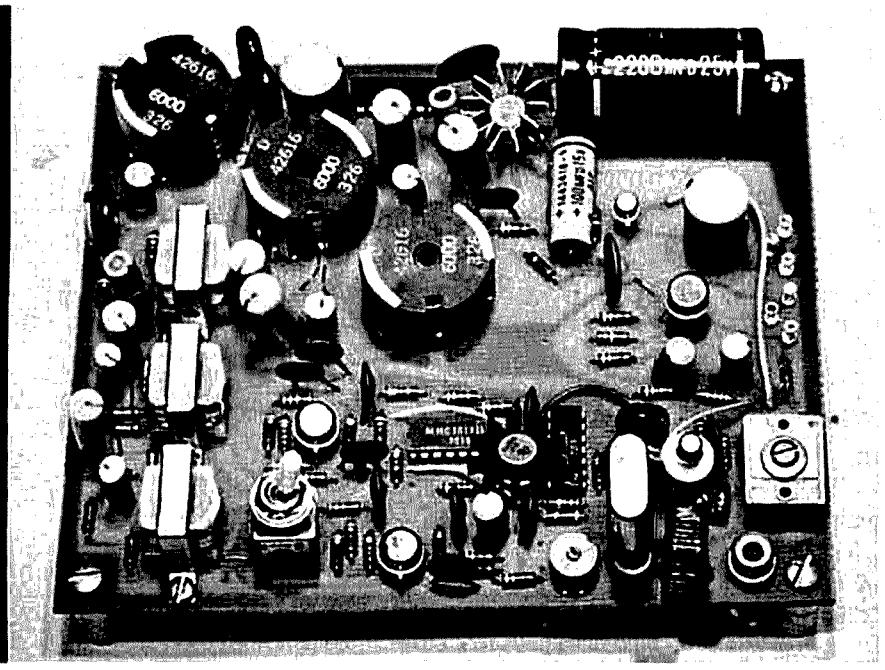
Since the proposed use of a portion of the 220-225 MHz band for CB would not be in accord with the International Table of Frequency Allocations, objections from Canada and Mexico might require a prohibition against any other operations in some border areas. Until that matter is resolved, mobile stations would not be permitted to operate within ten miles of the border, and base stations would not be permitted within 25 miles of the border.

Although I'm usually an optimist, there appears to be little that amateurs can do to prevent the loss of 224-225 MHz. When a proposal reaches the FCC Docket stage, it is usually approved in one form or another. If you remember the big-bucks EIA steamroller and the nearly one-million CBers clamoring for approval, the new class-E service is practically a foregone conclusion.

However, there's a small, bright side to the picture, too. Since the industry will be geared up to mass produce 225-MHz CB radios, there should be an abundance of gear available for amateur use. 220 has always been one of our little-used bands, essentially an orphan, and the lack of proper equipment has been one of the major stumbling blocks. On the other hand, it would be a simple matter for a proficient CB technician to illegally tweak his type-accepted radio down to the amateur frequencies. Perhaps that's the price of progress.

Never-the-less, the FCC wants to hear your comments. Don't forget the 14 copies.

Jim Fisk, W1DTY
editor



solid-state phasing-type ssb communications receiver

The Phase II Receiver — a state-of-the-art phasing-type receiver for high-frequency ssb communications

G.K. Shubert, WAØJYK, 1308 Leeview Drive, Olathe, Kansas 66061

Many amateurs participate in a public service net or a public service program such as MARS, CAP, AREC or RACES. This receiver, which I call the *Phase II Receiver,* was designed to provide economical monitoring of a single net frequency while the station receiver, or transceiver, is free for general hamming. Portable and mobile operation was also considered. The power requirement is 10 to 24 volts, ac or dc, at about 180 mA. Although this is a bit high for flashlight batteries, the design concept required ICs that draw relatively high current.

Crystal-controlled operation provides excellent stability and, since fixed frequency monitoring with high stability was the object, no vfo or frequency synthesizer was included in the initial design. However, it may be added at a future date. Most commercial equipment available to amateurs does not provide the frequency coverage or the stability required for use on MARS or CAP frequencies. It's really great to turn on a cold receiver and be within 10 hertz of the exact frequency — immediately! That is what the *Phase II Receiver* will do.

Direct conversion has been used often for inexpensive receivers but the highly undesireable audio image has been a menace, particularly on crowded bands.[1,2] Although many of these receivers have been referred to as ssb receivers, that is a complete misnomer. More accurately, they are double-sideband receivers that will copy single-sideband signals providing, of course, that there is no strong interference on the unwanted sideband. Even if there is absolutely no interference there is still a degradation of the signal-to-noise ratio because the receiver bandwidth is twice as wide as the signal.

phasing-type receivers

The phasing-type receiver or, at least, the phasing-type detector is by no means a new technique. The phasing-type transmitter, however, has retained its popularity better than the phasing-type receiver. Excellent explanations of the technique were given by Villard, Thompson and Norgaard in several post-war issues of *QST*.[3,4] Amateur ssb was in its infancy when those articles were written, and the phasing method of receiving and transmitting was at least as popular as the filtering method — probably more so because it held an economic advantage. Perhaps it still does!

Taylor revived the phasing receiving technique in 1969 and combined phasing with direct conversion.[5] For those of you with less extensive libraries and for those of you who understand semiconductors better than tubes, the phasing technique and direct conversion receivers were brought up to date by Ed Noll in 1971.[6]

It has been stated that a direct-conversion receiver does not provide single-signal reception.[7] This is somewhat like saying that a Ford automobile has one-hundred horsepower — a rather senseless generalization. There are quite a few filter-type receivers around that really don't provide single-signal reception *either*, due to design shortcomings.

I have presented the *Phase II Receiver* and the phasing technique of receiving at several club meetings and at two Regional Air Force MARS Conferences. It is interesting that the newcomers understand ICs, but some of the old-timers have a much better comprehension of the phasing technique.

To be competitive with receivers that use crystal or mechanical filters, the phasing-type receiver must suppress the unwanted sideband by 40 dB or more. Wade has researched the problem of unwanted sideband suppression thoroughly and provided the data in convenient form.[8] At first glance, the infinitesimal phase and amplitude variations that can be tolerated seem formidable but, modern integrated circuits maintain the amplitude balance, a digital rf phase shifter can be quite accurate, and audio phase shift can be controlled very closely with modern computer-aided designs.

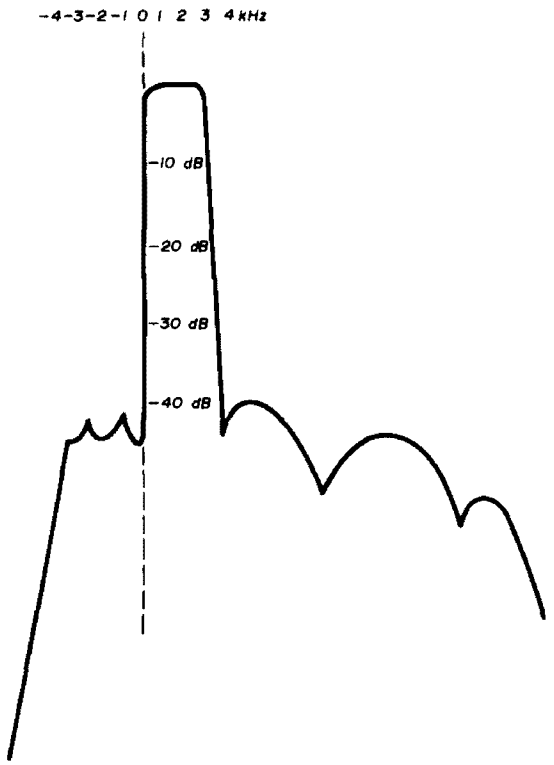


fig. 1. Passband characteristic of receiver for upper sideband reception.

selectivity

The passband shape of a phasing-type receiver is unique. Fig. 1 shows the receiver passband for upper sideband re-

ception. Cancellation of the lower sideband occurs immediately below the suppressed-carrier frequency and the passband shape on the upper side is controlled by the audio-frequency lowpass filter. The effective passband shape factor is better than 1.2:1 at the 40-dB points. A crystal filter with skirts that steep should easily cost more than the whole *Phase II Receiver.* A bandpass filter may be substituted for the audio lowpass filter for the reception of CW or RTTY.

Taylor stated that reception with a direct-conversion receiver gives the signal a feeling of "transparency" or "presence."[5] Impulse noise and atmospheric noise are less objectionable too. Crystal filters and mechanical filters have a very narrow percentage bandwidth with an accompanying high degree of delay distortion. Short pulses applied to this type of filter are severely distorted or stretched. A splendid example of this is to listen to a pulse with a wide filter (a-m type filter) and then listen to the same pulse with a very narrow filter (CW type). A pulse that only causes a click with the wide filter will result in a dull thud with the narrow filter.

The delay distortion of the audio low-pass filter, including the all-pass phase-shifter, is only a small fraction of that found in small percentage bandwidth bandpass filters. The all-pass networks used in the audio phase-shift networks provide a certain amount of time delay of the signal. The signal at this point in the receiver has quite wide bandwidth. Undoubtedly, these conditions will pave the way for the future addition of a digital noise blanker.

A design goal was to provide a versatile receiver module that could function by itself or be integrated into a monitoring or transceiving system. Since the system approach was in mind, very little front-end filtering has been provided on the PC board itself. The board has been drawn to allow the mounting of volume and agc potentiometers directly on the board, but in most cases panel-mounted pots would probably be more convenient. The complete *Phase II Receiver* is on a 5x6-inch single-sided PC board which is potentially the most valuable 30 square inches of real-estate in my hamshack.*

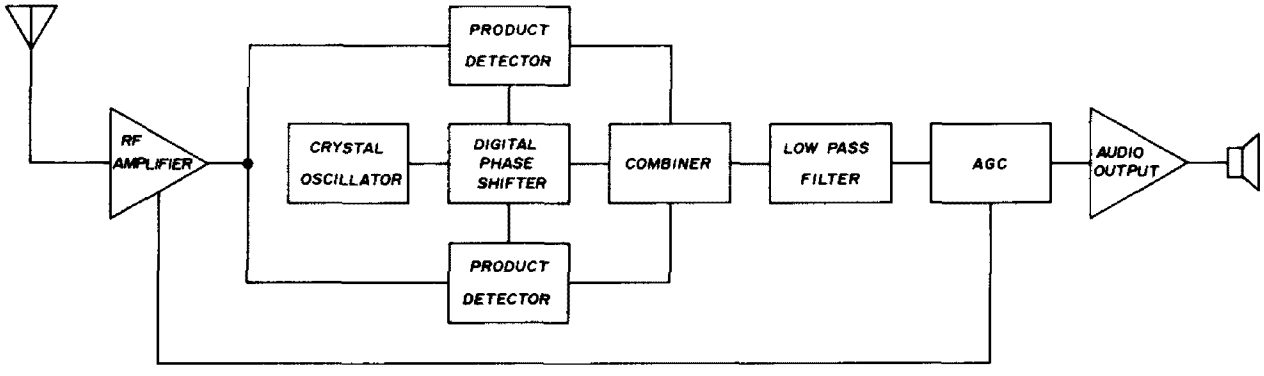


fig. 2. Block diagram of the high-performance phasing-type ssb receiver. The circuit, fig. 3, makes maximum use of integrated circuits.

Primary frequency coverage of the receiver as shown is from 3 to 10 MHz but this range can be extended by changing T1 or by providing a different front-end filter. Sideband suppression is poorer above 10 MHz and reception below 1 MHz may require more inductance at L1.

It is also possible to use the basic receiver for reception of vlf and low-frequency CW and RTTY transmissions. Of course, a-m reception is not very pleasant with no way to turn off the bfo,

*Printed-circuit boards for the Phase II Receiver, drilled and silver plated, are priced at $8.00 each from CMS Products, 317 NW 82nd Street, Kansas City, Missouri 64118.

so broadcast reception doesn't offer much. Frequency measurement of a received signal is a simple matter with this receiver and more will be mentioned about this later.

rf amplifier

A dual-gate mosfet provides good agc control and amplification over a wide frequency range with no neutralization. The mosfet provides wide dynamic range and low cross-modulation but also requires "backward" agc voltages. The agc voltage decreases as received signal strength increases. The single tuned circuit in the front end is an absolute minimum. This single circuit may be adequate for portable operation but operating in a metropolitan area will require some additional filtering to keep out broadcast stations and other local amateurs.

The output stage of a companion-transmitter may provide enough additional filtering. If the receiver is to be used with a transmitter of fairly high power level it might be wise to place protective diodes across the front end circuit. The Motorola MFE3006 transistor is the only one I have tried in the front end but others should work as well or better Some other transistors might be a little more rugged but my receiver hasn't blown one, even with 100 watts radiating from an adjacent dipole.

product detectors

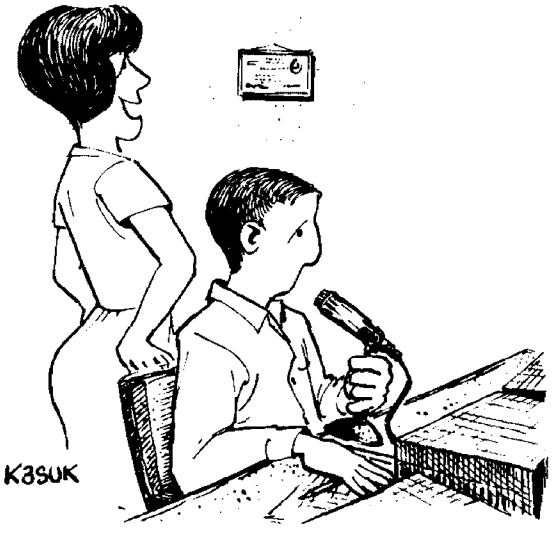
The signal from the rf amplifier is fed to a pair of identical MC1496 double-balanced mixer ICs used as product detectors. These product detectors have an extremely wide dynamic range and contribute quite a lot to the overall performance of the receiver. Transformer coupling at the outputs of the product detectors provides a low impedance to drive the audio phase-shift networks, attenuating frequencies below about 100 Hz and above 20 kHz or so. The frames of the transformers are grounded and help prevent any rf leakage from the product detectors to the audio combiner.

audio phase-shift networks

The audio phase-shift networks are a pair of four-pole, all-pass, RC networks which track each other with a very precise 90-degree phase difference.[9] The audio phase shifters in most block diagrams are shown as plus and minus 45 degree phase shifts but, of course, this is never the case in practical networks. It is extremely important to use precision film resistors in these networks to maintain proper phase shift not only initially but to stay put for the years to come.

If portable or mobile operation is contemplated the capacitors in this section should be polycarbonate film; for hamshack use the polyester film types (Mylar) should be adequate. Capacitor tolerance is equally important as the resistor tolerance but 1% capacitors are not a stock item. Even 5% capacitors are a little hard to find. If possible, these capacitors should be bridged or measured but 5% values and a little luck will provide good sideband suppression.

Two jfets provide the audio combining and some amplification with a pot connected between the sources to trim any amplitude imbalance preceeding them. The audio phase-shift networks prefer to operate from a zero impedance into an infinite impedance and are not too forgiving of any other conditions. The audio



"Hey! How about that. . . your license expired in March of 1969. . ."

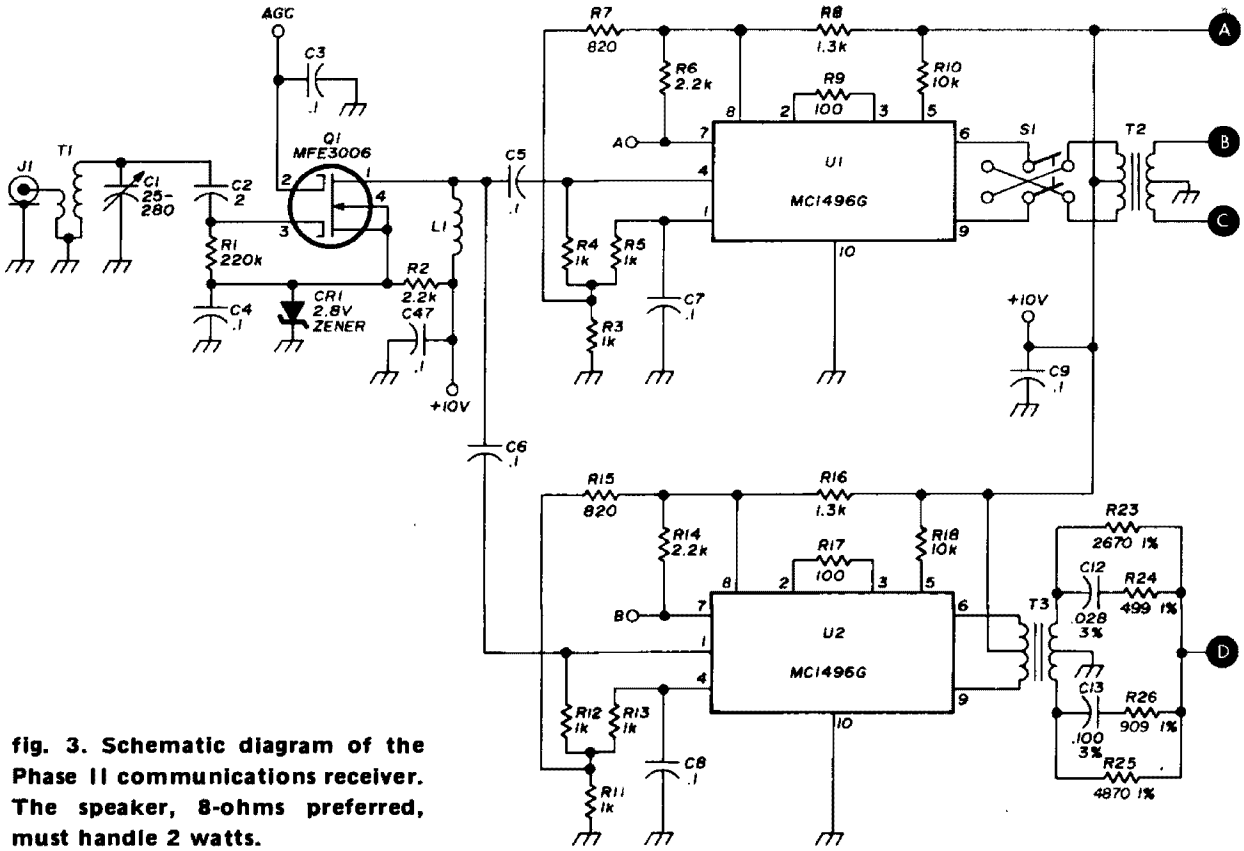
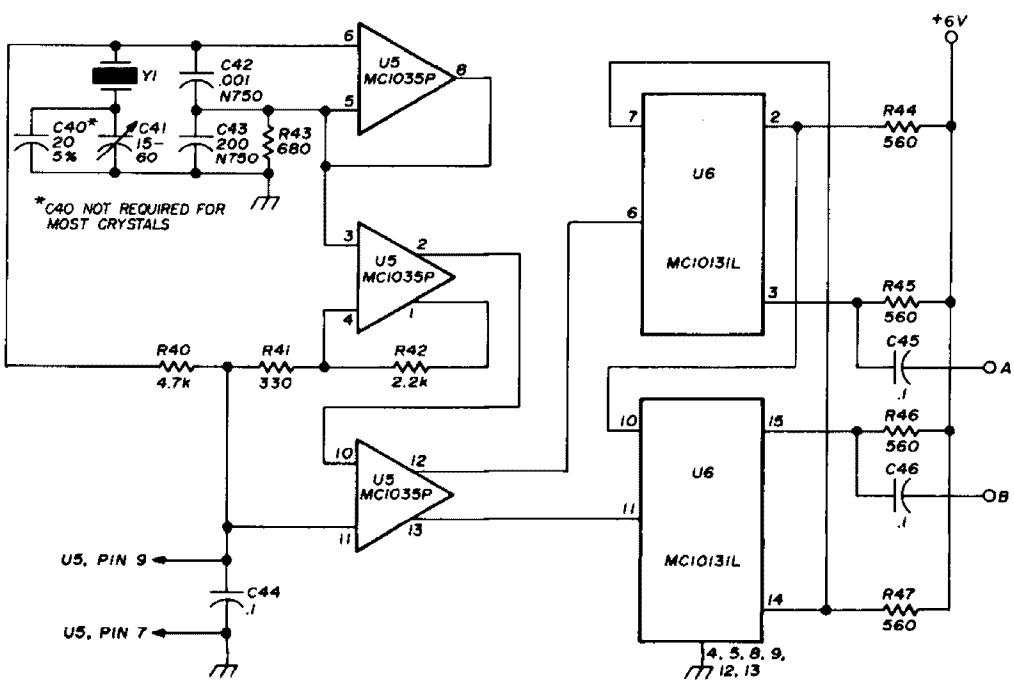


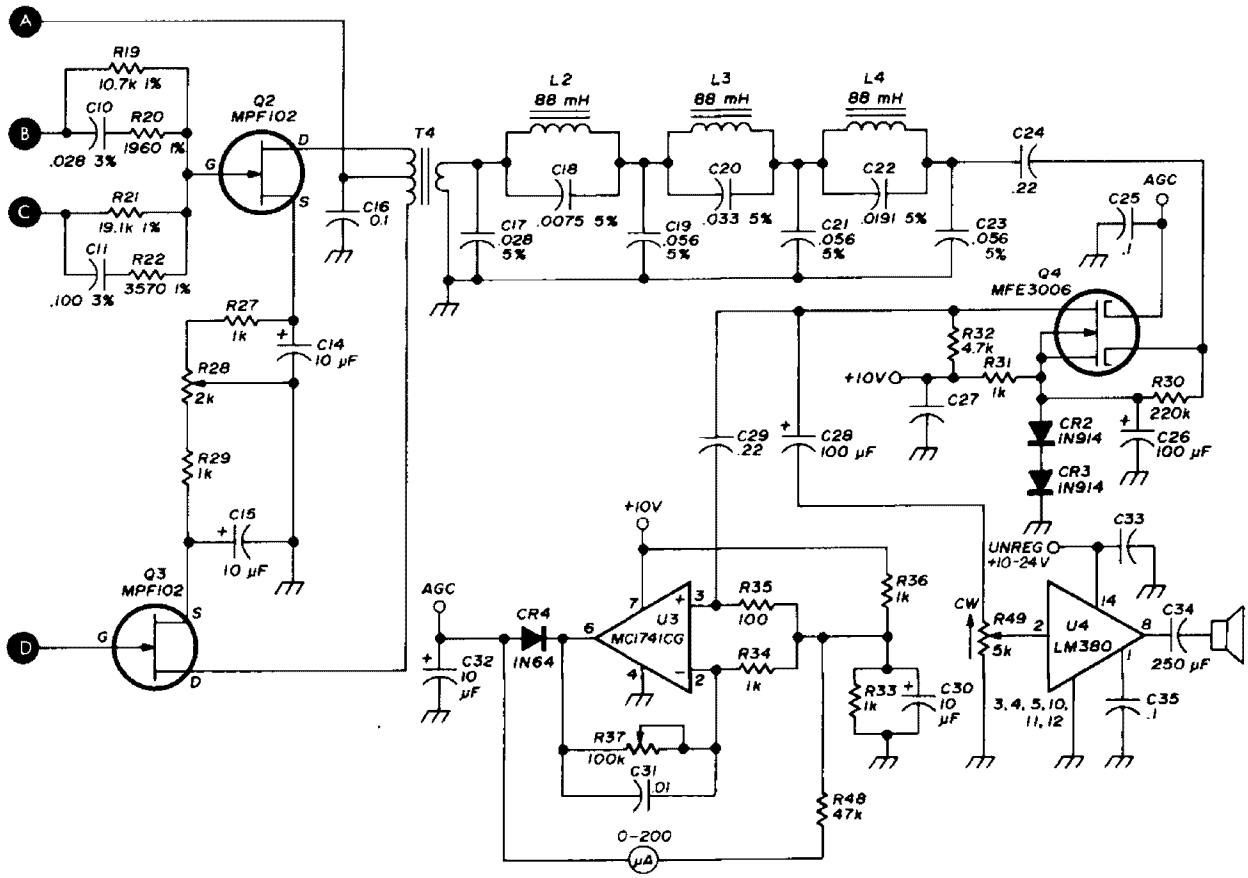
fig. 3. Schematic diagram of the Phase II communications receiver. The speaker, 8-ohms preferred, must handle 2 watts.

combiner uses a transformer for mixing and for impedance conversion to match the 1 kilohm input impedance of the audio lowpass filter.

audio lowpass filter

The audio lowpass filter is an elliptic function type design providing three attenuation peaks in the stopband and a cutoff frequency of 2.8 kHz. The filter has been computer "juggled" to permit the use of surplus 88-mH loading coils while retaining the best possible performance. Several receivers I have built used





L1	rf choke, 40 turns no. 34 on C40304TC core
L2	88-mH toroid, tuned to 6180 Hz with C18
L3	88-mH toroid, tuned to 3340 Hz with C20
L4	88-mH toroid, tuned to 3815 with C22
R28	2000-ohm pot, PC mounting (Mallory MTC-23L1)
R37	100k pot, PC mounting (Mallory MTC-15L1)
R49	5000-ohm pot, PC mounting (Mallory MTC-53L1)
S1	PC-mounting dpdt switch (ALCO MSTA-206N)
T1	secondary, 40 turns no. 24; primary, 2 turns no. 24 on Amidon T68-6 core (for 3 - 10 MHz)
T2,T3,T4	3:1 miniature audio transformer*
Y1	fundamental crystal operated at twice desired frequency

ferrite pot cores (Ferroxcube 2616 A1 400 3b9) but the slight bit of improvement is not worth the additional cost. The PC board is designed to accept either the toroidal loading coils or the pot-core assemblies. A 10% variation in the value of the capacitors will not make any noticeable difference to the ear.

This filter determines the passband shape and stopband performance of the receiver to a large degree. For those hardy souls who "roll their own" and wish to tailor a filter to suit some particular need, the three articles by Wetherhold will help lead the way.[10,11,12]

The input impedance to the filter is nominally 1000 ohms and the load is nearly 1 megohm. Active filters have been tried but, in this degree of complexity, are far too expensive, consume power, and are tempermental. Active filters generally require both positive and negative power supplies but the National LM 3900 quad-op amp IC may provide some relief for this problem.

Bandpass filters may be used for reception of CW and RTTY and some modification of the agc constants might be desirable too. It is also possible to derive afc voltage from the RTTY term-

*The set of three transformers is available from the author for $10.00, postpaid.

inal unit and feed it back to a voltage-variable capacitor across the crystal. Why not integrate the receiver right into the terminal unit? This is very handy for autostart operation.

audio-derived agc

Audio-derived agc is about the only means of agc available in a direct-conversion receiver. As mentioned earlier, the agc voltage is reversed because the agc control elements are two mosfets. One in the rf amplifier and another following the lowpass filter in the audio section. A 741-type operational amplifier provides the amplification for the agc independently of most of the speaker audio amplification. The response of this amplifier is modified with a feedback capacitor to slightly delay the attack time and prevent agc hangup on noise spikes.

The coupling capacitor to the agc amplifier is small to prevent response to very low frequencies. The information to be copied by an ssb receiver is a human voice and the object is to tailor the agc response to the voice and minimize response to undesired signals. A very broad peak at about 700 Hz is desirable. The amplifier pulls down the agc voltage quite rapidly and a resistor in series with the S-meter recharges a capacitor providing hand operation.

audio output amplifier

The audio output stage is one of the newer consumer products on the market, the National LM380 IC, 2-watt audio amplifier. This IC outperformed several other ICs and modules that were tried. It is very stable electrically, convenient mechanically, almost indestructable and, most important, inexpensive.

crystal oscillator and digital phase-shifter

Direct-conversion receivers of years past suffered from a lack of operational bandwidth. Some form of LC phase shifter covering a hundred kilohertz or so was considered broadband several years ago.[5] Digital phase shifters or quadrature generators are not new but this one requires only a frequency division by two rather than the more conventional division by four.[1,3]

A single inexpensive IC, a Motorola MC1035P, provides a crystal oscillator, a Schmitt trigger and a differential output amplifier.* The differential square-wave outputs from this stage are at the crystal frequency and, of course, 180 degrees out of phase. These square waves are divided by two in a switch-tail ring-counter which also provides synchronization to the phase quadrature, regardless of turn on or turn off of the receiver. One prototype receiver I built had no synchronizing circuitry and had a nasty habit of randomly choosing its own sideband.

This digital phase-shifter requires extremely fast, well matched ICs. The receivers built before the availability of the MC10131L just did not perform at 10 MHz with more than 20 dB of unwanted sideband suppression. Perhaps a faster IC will appear on the market which will permit operation to even high frequencies but the MC1496 ICs are also a limitation. Since it is nearly impossible to test the rf phase-shifter with conventional test equipment the performance of the overall receiver is the only practical test. The MC10131L IC is not cheap, and it draws a lot of current, but it is the key that unlocks the door to broadband operation.

The crystals are standard fundamental types for 32-pF operation. The PC board has both a trimmer and a shunt capacitor to permit a wide choice of shunt capacitance across the crystal. The crystal frequency is always twice the desired suppressed-carrier frequency. If a crystal is being ordered to copy CW or RTTY don't forget to choose your favorite sideband for copying plus a 1-kHz offset or so. Surplus CR-18/U crystals work fine as do

*Crystals are available from Buck-Man Electronics, 120 South Blake, Olathe, Kansas 66061. Specify crystals for the Phase II Receiver. 1 or 2 crystals, $7.00 each; 3 - 5 crystals, $6.00 each; 6 - 15 crystals, $5.00 each; 16 - 25 crystals, $4.00 each, postpaid.

many of the overtone crystals oscillating on their fundamental frequencies.

power supply

The receiver circuitry may be powered with ac or dc but if raw ac is used the audio level must be kept low or it will be extremely rough sounding. It is best to full-wave rectify the ac or provide a filter capacitor much larger than the 2000-μF used on the board. The 10-volt regulator also acts as an active filter to clean up the dc fed to all the low-level circuits. The second regulator provides 6.0 volts for the digital circuitry. Both regulator transistors should have clip-on type heat sinks since they may have to dissipate a couple of watts under extreme conditions.

construction

With some careful shopping it is possible to build the whole *Phase II Receiver* for about $60. A club or group purchasing at quantity prices can reduce the cost further and put life back into the club net, too. Most components are standard stock items and values have been selected to permit quantity purchases. The miniature transformers are not standard but are available below the cost of comparable standard American-made transformers.

The transformers have a red dot on the side with the most turns. The red dot on T2 and T3 is toward the product detectors and the dot on T4 faces R28, the sideband balance pot.

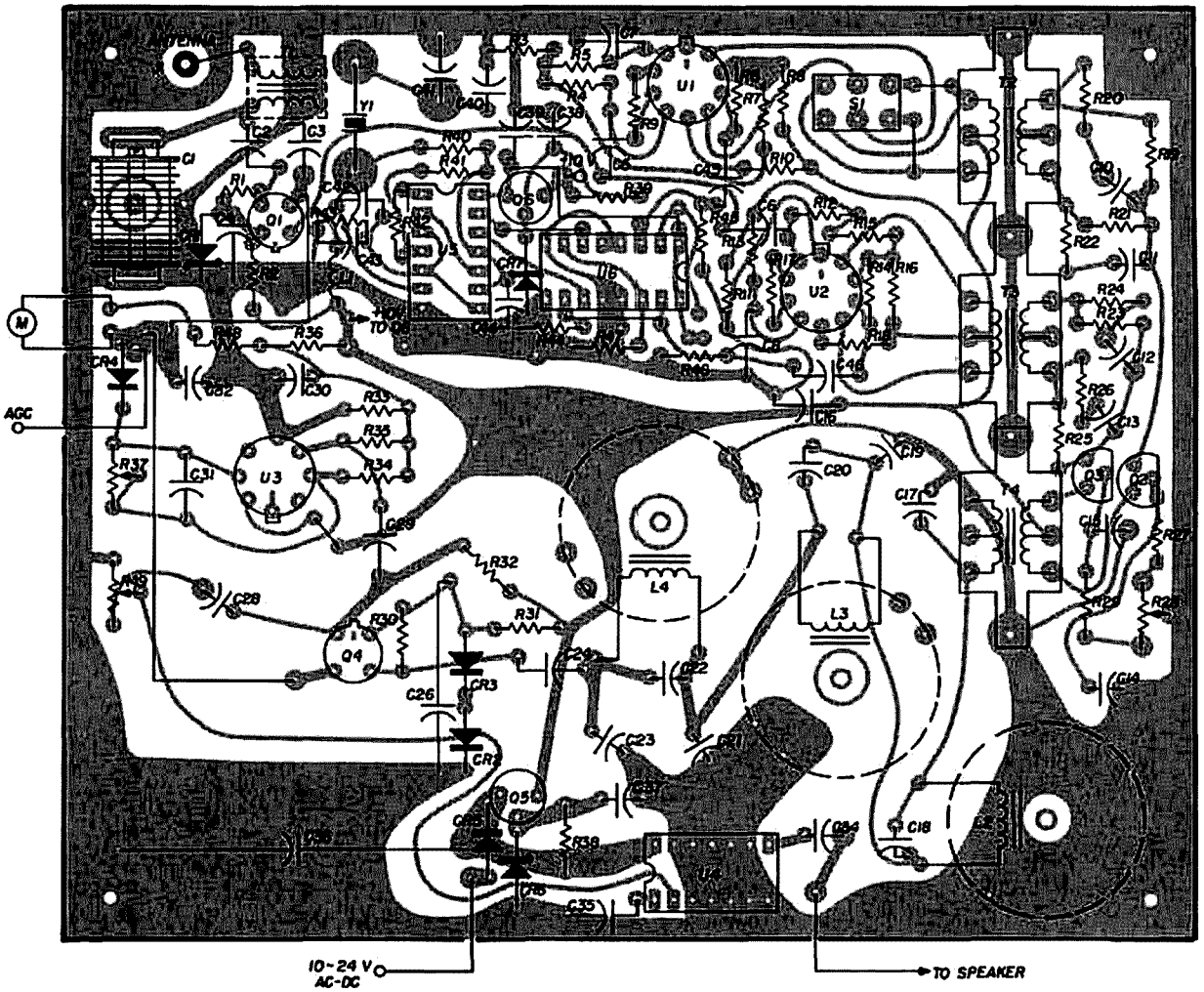


fig. 4. Printed-circuit layout of the complete phasing-type communications receiver.

S-meter

Sensitive meter movements are expensive and now even the surplus stock of 100-μA meters is about gone. A single inexpensive transistor and one resistor will permit use of a readily available 1-mA meter. Most of these meters are not calibrated in S-units or even in dB. The meter in my receiver is very appropriately calibrated in, "wind speed — miles per hour." On 3920 kHz the incoming signals generally run about 80 mph at mid-evening. The transistor and resistor may

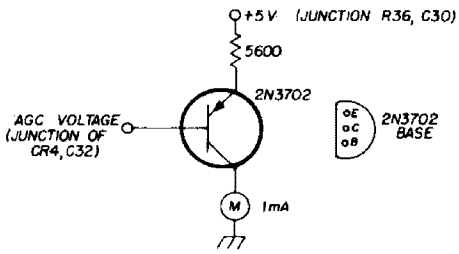


fig. 5. If sensitive 100-μA meter is not available for the S-meter, this transistor amplifier and 1-mA meter may be used.

be mounted on the board near the agc components or hung on the back of the meter (see fig. 5). Be sure to jumper the old meter pads together to maintain agc action.

Some of the cheap and surplus 741 IC op amps do not have as much output swing as they should and will not pull the agc voltage below 2.0 volts. Even with a good amplifier the voltage will not drop below 1.8 volts. This limits the dynamic range and large signal-handling ability of the receiver. The new Norton or current-mode op amps may be a solution to this problem as well as to the problems of splitting the power supply voltage and floating the amplifier. The LM3900 quad operational amplifier was not available during the development of this receiver but it might be incorporated into a future design with the other three amplifiers used for noise blanking.

alignment

The only adjustments to be made to the completed receiver are to balance the audio combiner pot for best sideband suppression and to adjust the front-end filter to peak the signal. Of course, the crystal frequency can be trimmed on the nose with the small trimmer capacitor in series with the crystal. Since the product detectors are driven with square waves it might be suspected that there would be strong spurious responses at the odd harmonics.

Data for the MC1496 double-balanced mixer indicates that the second and third harmonic sensitivity should be at least -30 dB even with extremely strong overdriving.[15] The carrier input to the product-detectors causes switching action anyway.[16]

troubleshooting

Troubleshooting the receiver is not difficult. The audio output stages can be used to self check the rest of the circuit back to the product detectors. Just remember that the agc voltage is reversed, about 5 volts for a weak signal and nearly 1 volt for strong signals.

A general-coverage receiver will provide a quick check on the crystal oscillator and the digital phase-shifter. You should receive a signal at the crystal frequency and at the desired receiving frequency at one-half the crystal frequency. Integrated circuits U5 and U6 are emitter-coupled logic and are intentionally operated above their normal 5.2-volt supply. The Phase II receiver's low-voltage supply is nominally 6.0 volts which is still safe for the ICs. The oscillator operates more reliably over a wider range and the phase-shifter is slightly more accurate at the higher supply voltage.

operation

Operation of the *Phase II Receiver* at voltages above 20 volts dc is not recommended because of the voltage limitation of the LM380 audio power amplifier IC and the wattage rating of R38. Transistor Q5 gets a little hot under the collar, too. If continued operation at voltages above 20 volts is desired, a 10- or 12-volt, 10-watt zener diode may be added in

series with the power supply line. If only portable battery operation is desired, all the components in the first voltage regulator may be removed since they will only consume power and regulate nothing. However, the input diode, CR5, should be left as a polarity reversal protector.

Unwanted sideband suppression is improved by adding a capacitor from one side of the primary of transformer T2 to the center-tap. This capacitor value varies from 2000 to 10,000 pF with different receivers, but in all cases makes some

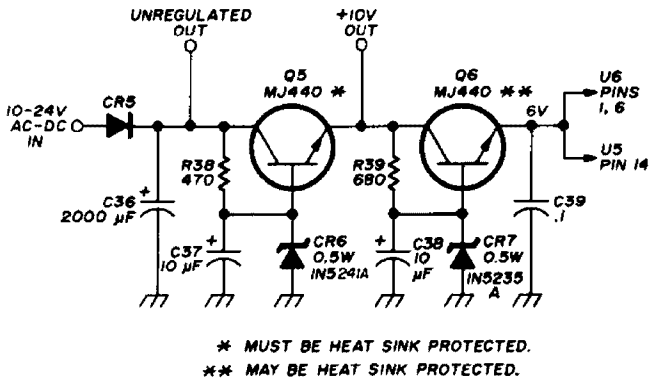


fig. 6. Power supply for the phasing-type receiver operates from 10 to 24 volts, ac or dc.

improvement. Circuit pads have been included on the PC board for this purpose.

Direct-conversion receivers have a peculiar ability that occasionally causes problems. The receiver can receive its own local oscillator. Logically, it would seem that this signal leakage out of the digital phase-shifter would always be within 90 degrees of the received frequency and would be a dc voltage coming from the product detectors. Transformer coupling was used in an attempt to minimize low frequency audio noise caused by self-reception. Certain conditions aggravate the self-reception problem, resulting in an obnoxious buzz. To minimize this problem the following procedures may be helpful. They are listed in order of decreasing importance.

1. Disconnect the frequency counter from the receiver when not taking actual frequency measurements.

2. Ground the receiver to a good ground.

3. Locate the antenna as remotely as possible and use coax feed.

4. Place the receiver board in a well shielded enclosure.

Using a converter with the Phase II receiver seems to present no problems as long as coax is used for signal input.

Strong signal handling, cross-modulation, and overload may be improved by increasing the voltage of zener diode CR1. Although this will result in some loss of sensitivity, on most frequencies below 10 MHz extreme sensitivity isn't all that useful.

No muting feature was provided since it was assumed that the audio output would simply be turned off if a transmitter was in operation. Experimenting with the companion breadboarded phasing transmitter has turned up two more possibilities. Since it is possible to receive your own signal while transmitting it may be desirable to pull the agc line down near ground potential and use the receiver audio as a sidetone signal. The audio can be muted by grounding pin 1 of the power amplifier, U4. A combination of both of these muting methods may be used but the agc line should be left slightly above ground to speed recovery time when switched back to the receive mode.

frequency measurement

Off-the-air frequency measurements are quite easy because there are two unused outputs from the digital rf phase-shifter, either of which may be fed directly to a frequency counter. The exact suppressed-carrier frequency of a ssb signal is quickly read.

Reading the frequency of a CW signal or a continuous carrier is much more difficult, believe it or not. Zero beating will not work because there is no output at frequencies near zero beat. The best method I have found for measuring the frequency of CW signals is to compare the 1-kHz output of the counter with a 1-kHz beat note from the receiver on a scope.

This provides exactly 1-kHz offset and can be added to or subtracted from the counter reading, depending upon the sideband used. It is possible to preset some counters to allow for the 1-kHz offset.[14] Crystal drift experienced at 7.6-MHz was only plus or minus 5 Hz over a period of several months.

summary

Four prototypes have been in service for over a year. No two of the prototypes are exactly the same but they all have performed quite well with the exception, perhaps, of the one that chooses its own sideband, but that problem has been cured. Frontend filters have been an absolute necessity to permit operation of adjacent transmitting equipment. Performance is good enough that the Collins 75A-4 has been on the shelf for six months and the R390-A is a standby.

The most obvious follow up project for the receiver is a companion transmitter and a vfo. A breadboard transmitter has been in operation for several months with excellent results, considering the output power of one watt. Signal quality has been the key to success with such low power.

The transmitter is a sort of *reciter* because it does transceive with the receiver and it uses the receiver rf digital phase-shifter. It provides a couple of unique features: it is possible to transmit on one sideband and receive on the opposite — even simultaneously, within the limits of the unwanted sideband suppression. It is also possible to transceive but to monitor your own signal off the air would require a separate receiver and transmitter.

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ham radio



"This is incredible. . . I've contacted a Captain Lightning in his Mars space ship."

high-gain log-periodic antenna for 10, 15 and 20

This 17-element log-periodic antenna provides approximately 13-dB gain over three amateur bands

George Smith, W4AEO, 1816 Brevard Place, Camden, South Carolina 29020

In a previous article I described two fixed, doublet-type log-periodic antennas which cover the entire frequency range from 14 to 30 MHz.[1] Although both of these antennas have 12 elements, one, 70-feet long, provides approximately 10-dB gain. The other, 40-feet long, gives approximately 8-dB forward gain.

Since writing that article I have built eight more log-periodic antennas for various frequency ranges, including several for 40 meters.[2] One of these log periodics which is especially interesting uses 17 elements to provide 12- to 13-dB forward gain on the amateur 20-, 15- and 10-meter bands. This antenna is of interest to amateurs who want a single fixed wire beam aimed in one particular direction and it will be described here.

the antenna

The high-gain, three-band log periodic shown in fig. 1 is 100-feet long and has an apex angle of 16 degrees ($\alpha = 8^\circ$). The minimum space required for this antenna is approximately 125-feet long by 80-feet wide. Four masts or other supports are required. The antenna is suspended between the four masts with two nylon catenary side lines as shown in fig. 2.

Unlike the antenna described in reference 1, the center open-wire feedline carries some of the weight of the antenna. This eliminates two of the nylon support lines used in the earlier design.

I used number-15 aluminum wire for both the center feedline and for the 17 active elements. The use of aluminum wire minimizes overall weight of the large antenna. This wire is manufactured for use in electric fences and is available from Sears (catalog number 13K22065) at $8.70 for a quarter-mile roll. Approximately 570-feet of wire are used in this three-band log-periodic antenna.

Although this aluminum electric-fence wire is strong and inexpensive, be careful not to put any kinks in it when you are putting the antenna together – aluminum wire breaks easily after it has been once kinked. As with any aluminum wire, special precautions must be taken when making electrical joints and splices.

If you use sturdy steel masts or wooden poles for the four antenna supports, copper wire can be used for the antenna.

However, the weight of the antenna will be much greater. At my station I use 75-foot-high pine and cedar trees for supports, so *antenna* weight must be kept to an absolute minimum.

construction

As with my previous log periodics, the center insulators are made from ¼-inch thick lucite or plexiglass as shown in fig. 3. The center element ends are attached to the two outer holes; the center open-wire feedline passes through the other two.

The center insulators are secured to the open-wire feedline by serving several turns of number-18 tinned copper wire over the feeders on each side of the insulator as shown in fig. 4.

The photograph shows a mockup of the center feedline and element arrangement. (In this mockup the elements are much more closely spaced than in the actual antenna.) Ceramic egg insulators are required at the front (element 17) and rear (element 1) because the homemade insulators will not take the stress at these points.

Ceramic egg insulators can also be used for the end insulators for the front and rear elements, up to 250 watts. For higher power, 6- to 8-inch Isolantite antenna insulators are better. Monofilament fishing line, 40- to 50-pound test, is used as end insulators for the rest of the elements. I used 40-pound test monofilament from Sears (catalog number 6KV32232), priced at $1.88 for a 325-yard spool.

The center feedline to each of the elements must be transposed as shown in fig. 1. The photograph shows that the necessary crossover is located below the center insulator — this provides a rain drip and lowers the center of gravity of the antenna. Since this log periodic uses an odd number of elements, the front and rear element connections do not have to be transposed, and the center feedline can be connected directly to the ends of the front and rear elements. You can use a 4:1 balun and coaxial line or 300-ohm tv twinlead with this antenna, as described in the previous articles.

The photo also illustrates the aluminum-to-aluminum connections between

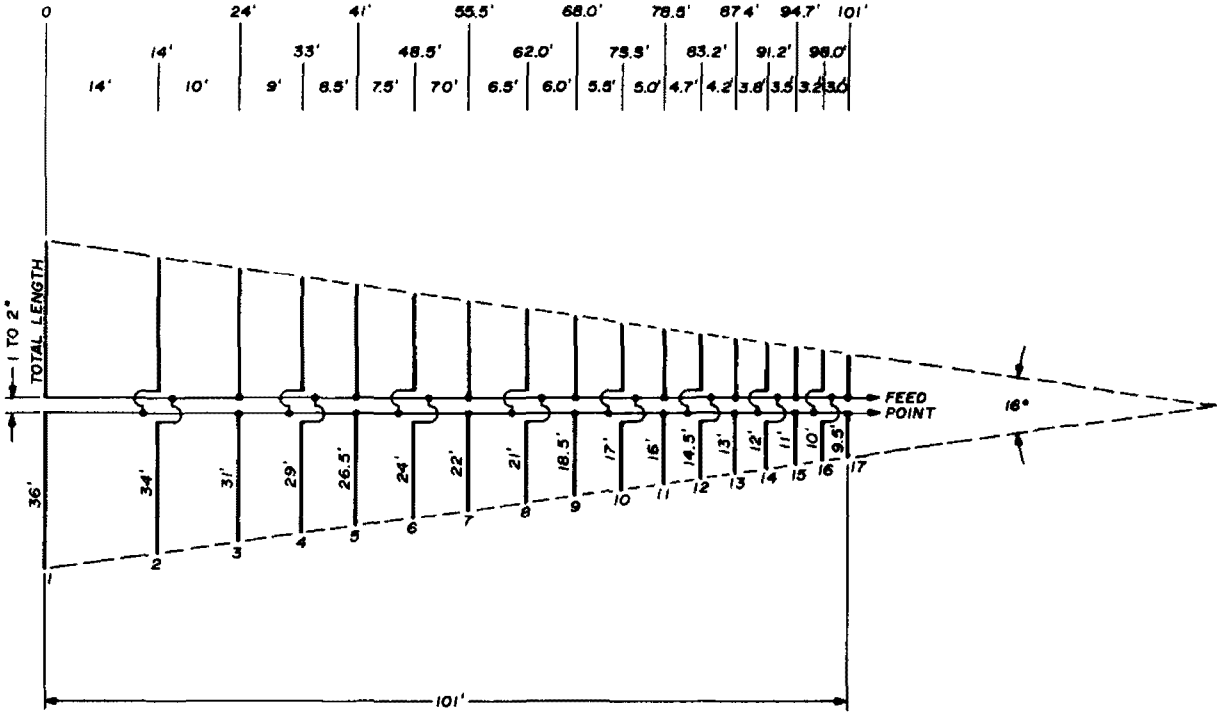


fig. 1. High-performance log-periodic antenna for 20, 15 and 10 meters uses 17 elements to provide approximately 13 dB gain. Installation requires four support masts and a space 125-feet long by 80-feet wide.

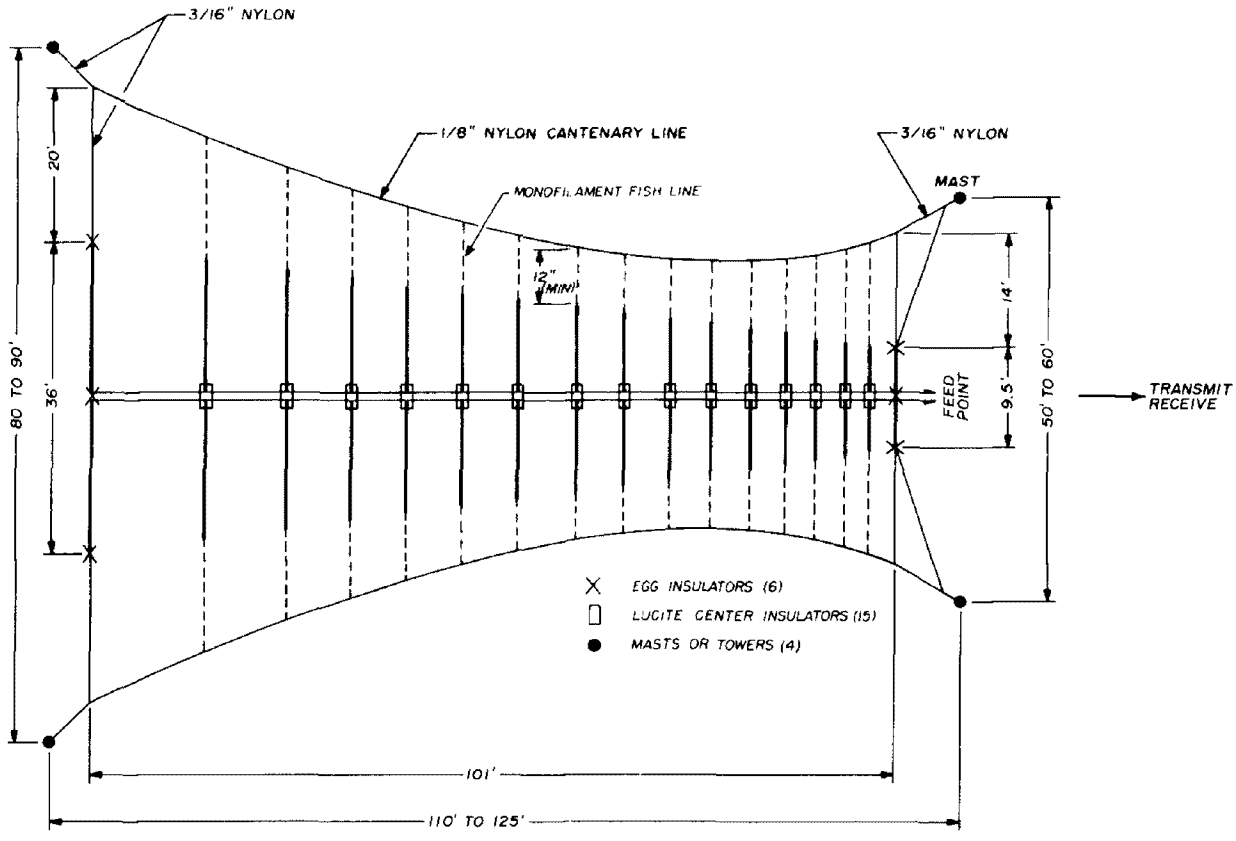


fig. 2. Installation of the 17-element log-periodic antenna. Masts should be a minimum of 40-feet high — 70 feet is better.

the elements and the center feedline. The center element ends are cut 10 to 12 inches longer than the specified element length. This extra length is then fed through the end hole of the center insulator and given three wraps over the element to secure it to the insulator. The remaining length is connected to the center feedline.

All aluminum-to-aluminum electrical connections use mechanical joints. When connecting the transmission line or balun to the antenna, use stainless steel or cadmium-plated hardware. By using large cadmium-plated shakeproof washers with internal teeth good contact can be made to the aluminum wire. I used this method with my first log periodic, and it has been in constant service since September, 1970, with no corrosion or electrolytic problem, to date.

If copper conductors must be used at the feedpoint, use large cadmium-plated terminals or solder lugs, bolting them to the aluminum wire with plated screws and nuts. This eliminates any direct aluminum-to-copper contact that could lead to problems with electrolysis.

performance

This is the highest gain log periodic I have built so far. It is beamed west and is installed about 60 feet above the ground. Stations in California often tell me I am the strongest W4 on the band. Considering that I'm using a barefoot transceiver, while most of the competition are running 1000 watts, it gives me satisfaction to know that by properly using a few hundred feet of wire, such gain is possible for less than $20.00.

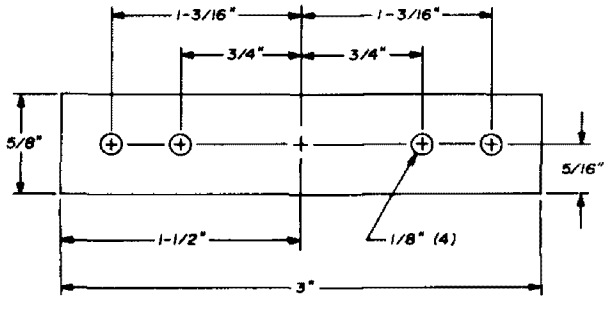
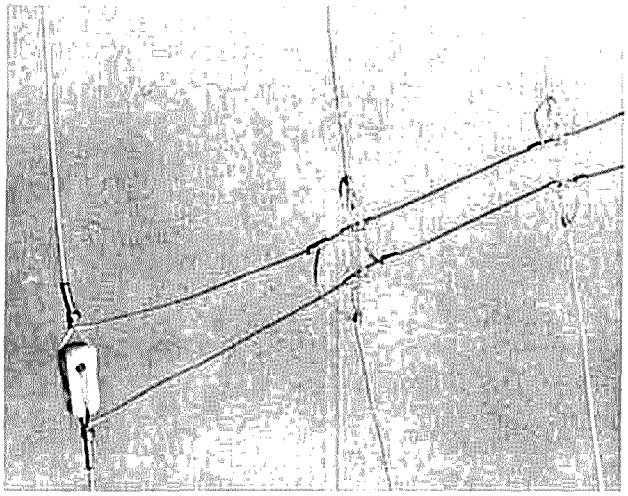


fig. 3. Center insulators for the log-periodic antenna. Material is ¼" lucite or plexiglass.



Mockup of center feedline showing installation of the center insulators and mechanical aluminum-to-aluminum connections.

I have not tried to make quantitative front-to-back or side attenuation tests, but from on-the-air tests signals from stations off the side of the antenna seem to be about 30-dB down. This is a great help here for attenuating the extremely strong signals that pour in from South and Central America.

I have not been able to obtain much data on the front-to-back ratio as there is not much to the east of here except the Atlantic Ocean. However, I was recently monitoring a VK1 in the evening by long path from the east. He was about S4 on my non-gain, non-directional antennas but was absolutely nil off the rear of the 17-element log periodic. I was also recently monitoring a CR6 in Angola – he was good S5 copy with my non-gain antenna, but was only S1 or less off the back of the log periodic. These simple observations seem to indicate that this log periodic has a better front-to-back ratio than the other log periodics I have built in the past.

When I designed this long log periodic I tried to improve the front-to-back ratio by increasing the spacing between elements 1 and 2 to 0.2-wavelength. This is 14 feet, as opposed to a spacing of 11 feet which would be normal. Normally the front-to-back ratio of a log periodic is 14 to 15 dB minimum. However, to date I have not been able to determine if the increased spacing between elements 1 and 2 was of any help. The rest of the element spacings are normal for a log periodic having this boom length and apex angle.

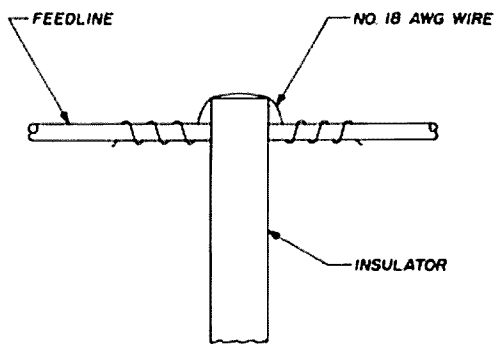


fig. 4. Lucite insulators are secured to the center feedline with a few twists of no. 18 wire.

summary

If you need a high-gain, fixed beam and have the available space, a long log periodic should be considered. Even a long log periodic of this type requires less acreage than a rhombic antenna designed for the same gain on 20 meters. Furthermore, unlike the rhombic, the gain of a log periodic does not fall off at the lower end of the frequency range. If anything, the log periodic seems to perform slightly better at its low frequency end.

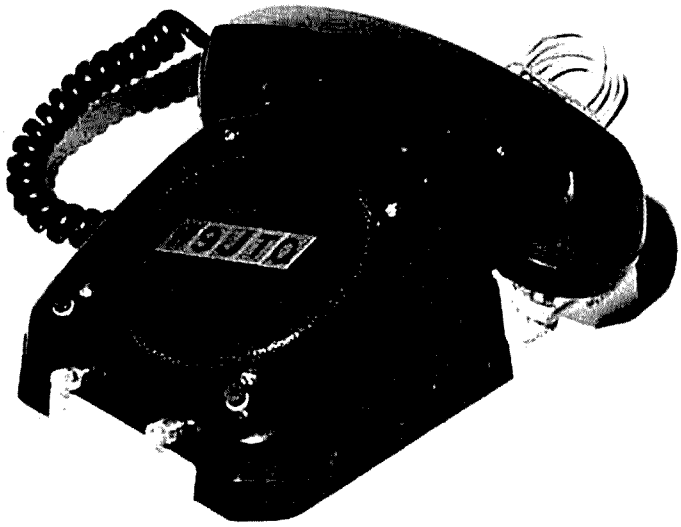
table 1. Price list of materials for the 17-element log-periodic antenna.

600-feet	no. 15 aluminum electric-fence wire	$ 3.60
250-feet	1/8" nylon line	7.50
100-feet	3/16" nylon line	5.00
25 sq. in.	1/4" lucite for center insulators	1.00
6 each	ceramic egg insulators	.90
1 roll	40-pound monofilament fishing line	1.88
		$19.88

references

1. George Smith, W4AEO, "Three-Band Log Periodic for 20, 15 and 10," *ham radio,* September, 1972, page 28.
2. George Smith, W4AEO, "40-Meter Log Periodics," *ham radio,* May, 1973, page 16.

ham radio



two-meter fm base station

A compact rig with 12 channels built into a telephone desk set

Al Ericson, W9JTQ, South Washtenaw Avenue, Chicago, Illinois

For the amateur who wants his 2-meter fm base station in the living room, this desk set should be the answer. At a glance it appears to be just another telephone. But underneath are 12 channels for transmitting and receiving, each individually selectable. The desk set uses Motorola HT-220 handy-talky circuit boards, which are easily adaptable to a compact rig such as this. In addition to the rf circuits, an ac regulated power supply completes the desk set base station.

If you're close to a repeater or have an efficient antenna, the two watts output from the desk set will be sufficient to capture the repeater with full quieting. More power can be obtained by adding an outboard solid-state amplifier, such as the Dycomm 101-500E, which can be remotely located and fed with coax cable. The speaker in the desk set, with the 1/2-watt audio output from the receiver section, is ample for base-station use.

The photos and schematic, together with the instructions furnished, should provide sufficient guidance for those who would like to duplicate the desk set.

construction

Space is limited inside the telephone desk set, but with care all components can be accommodated. A parts list is provided in table 1.

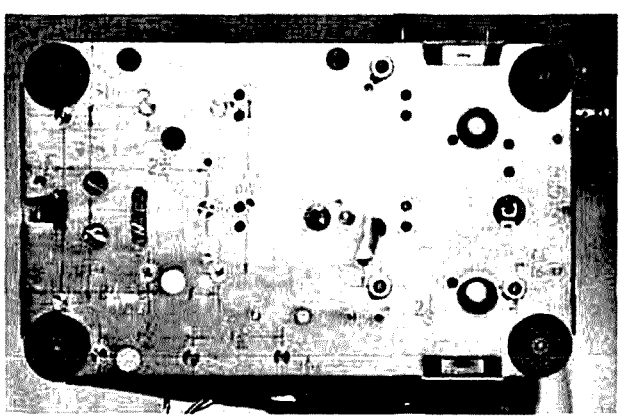
The following additional holes must be drilled in the desk set metal base (see photos for dimensions):

1. Three holes for mounting the HT-220 chassis assembly.

2. Three holes for the dc power-supply circuit board.

3. Two holes for the power-supply transistor heat sink.

4. Four holes for the power-transformer mounting bracket.



Baseplate hole layout.

5. Two holes and a small cutout in the rear lip of the base for the coax rf output fitting (not visible in the photo of the base).

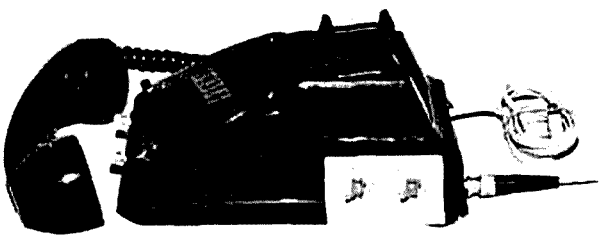
Drill and file the following holes in the plastic case:

1. Two holes for volume and squelch controls (front edge).

2. Two holes for channel selector switches (right side).

3. One slotted hole on rear edge for the coax connector.

4. Four holes for mounting the ac line switch and fuse holder.



Separate channel selector switches for transmit and receive.

The following parts and subassemblies are mounted on the telephone desk set:

1. HT-220 chassis assembly (refer to photos and description).

2. 110 volt/24 volt ac power transformer with two right-angle metal mounting brackets.

3. Dc power supply circuit board with three 3/8-inch-high mounting posts.

4. Dc power supply regulator transistor, mounted on sheet-metal heat sink.

5. Coax connector, mounted to rear edge of base.

6. Ac line switch, fuse, and line cord.

7. Volume and squelch controls and indicator lights.

8. 40-45 ohm dynamic speaker.

description of subassemblies

The HT-220 circuit board is mounted in a Motorola part No. 7D83124H-1-D metal base, which acts as the heat sink for the final amplifier transistor. This assembly is mounted on 5/16-inch posts at the four corners of the telephone desk-set

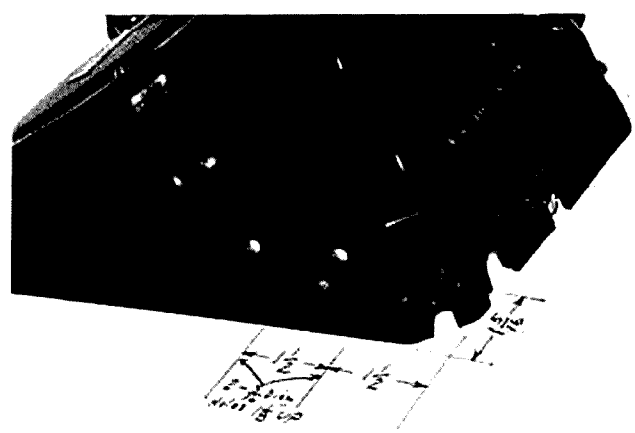
table 1. Complete parts list for the compact two-meter base station.

quantity	item
1	Motorola desk set housing with cradle switch, speaker grill, and handset
1	Motorola HT-220 circuit board with control relay for remote operation
1	Motorola 7D83124HO1-D metal chassis housing with mounting screws for HT-220 circuit board
4	3/16 in. diameter x 5/16 in. long spacers to mount housing to telephone base
1	Motorola 15E84004/HO1 plastic spacer sleeve with mounting screws for crystal deck circuit board
1	chassis-mount BNC coax connector for antenna output
1	25k volume control - use miniature control from HT-220 circuit board - part no. 18-83561HO2
2	5k squelch control (Motorola part no. 18C83560H01)
2	miniature knobs for above for 1/8 in. diameter shaft
2	12-position miniature Grayhill switches for channel selectors (part no. 51YY23047-1-12N)
2	bar knobs for above (for 1/8-in. shaft)
1	4-inch dynamic speaker with 40-50 ohm voice coil
1	miniature mike cartridge for Motorola HT-220 with rubber mounting pad (part no. 50-82157JO1)
2	24-volt Dialite pilot lights
1	110-volt/24V 1 amp transformer (Olson part no. T-290)
4	1-amp 200-volt rectifier diodes (Radio Shack part no. 276-1102)
2	500 μF, 35-volt electrolytic capacitors (Radio Shack part no. 272-1018)
1	Motorola HEP-247 transistor for regulator
1	Motorola HEP-607 15-volt 1/2-watt zener
1	1500-ohm 1/2-watt resistor
1	3/4-amp AGC fuse (Radio Shack part no. 270-1272)
1	3-3/4 in. x 2-5/8 in. perf board (punched) for power supply (Olson part no. K-116)
1	spst toggle switch (Olson part no. SW-384)
1	panel-mounted fuse holder (Radio Shack part no. 270-364)
1	6-terminal barrier strip (Radio Shack part no. 274-650)
1	special Motorola fuse (do not substitute) (Motorola part no. 65B82896B03)

Note: The Motorola HT-220 circuit boards, desk set, and Motorola parts are available from Spectronics, In., 1009 Garfield Street, Oak Park, Illinois 60304. When you order your HT-220 circuit board, request the schematics. Cost is 10 cents per 8-1/2 x 11 page, or about $1.00 a set.

base. The two Grayhill part no. 51YY23047-1-12N 12-position channel switches are mounted on the 1/8-inch-thick battery end of Motorola part no. 15E84004H01 plastic extension (see table 1) for the thick-pack HT-220, from which the battery section is sawed off and discarded. The hollow tubular member that houses the collapsible antenna must also be sawed out of the inner part of the plastic extension sleeve to provide space for the transmit and receive crystals, their coils, and the channel switches. Note that this sleeve is turned 180 degrees from the way it is normally mounted in a handy-talky.

A special circuit board was prepared for the 20 channels on the board which, including the four channels on the HT-220 board, gives 12 transmit and 12 receive channels. This board has holes above each tuning slug in the HT-220



Hole dimensions for channel selector switches and antenna coax connector.

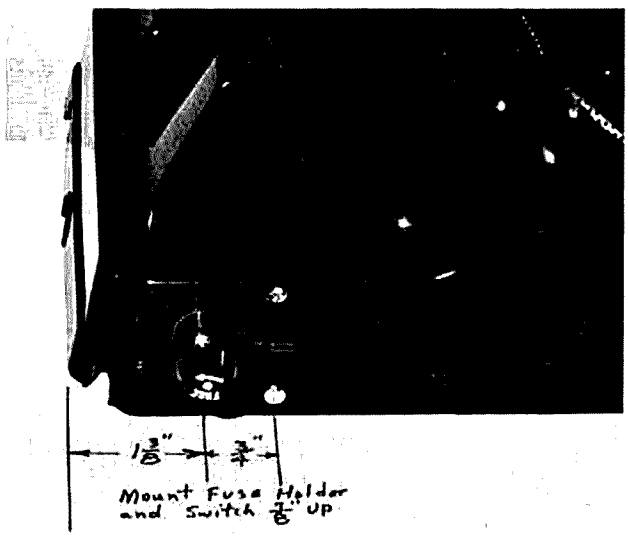
chassis, so that the rig can be tuned without disassembling the crystal deck from the HT-220 deck. One crystal coil slug is inaccessible from the top, but a hole already exists in the bottom of the desk-set base through which it can be adjusted.

A terminal strip is fastened to the side of the HT-220 chassis and is used to connect the telephone handset terminals to the circuits. Squelch and volume controls are located on the front skirt of the handset, so leads must be extended from the HT-220 chassis assembly to these controls, the dc power supply, and the handset cradle switch. See photos for details and refer to the schematic (fig. 1).

HT-220 board preparation

The HT-220 circuit board, as received, contains a relay for remote control of a speaker and microphone. Unsolder and remove the following:

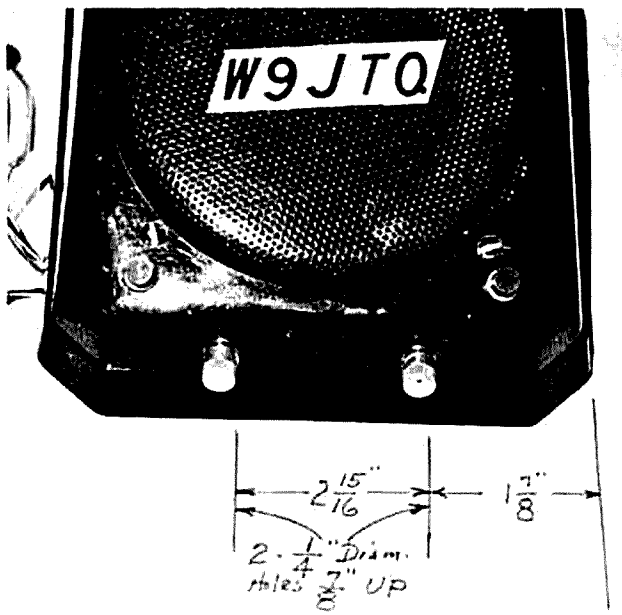
1. Green lead from pin 10 on IC 101 on the back side of the board.
2. Blue/green lead.
3. White/brown lead.
4. Brown/green lead.
5. 80-ohm resistor and green/white lead.
6. Ground lead (black) at edge of board by relay.



Mounting dimensions for ac power line fuse and switch.

Do not remove the following:

1. Grey, yellow, violet leads (squelch control).
2. White/orange, white/yellow, white/blue leads (receive crystal switch).
3. White/red, white/green, white/brown leads (transmit crystal switch).
4. Black, brown, green leads (volume control).
5. Black/red (relay-control lead).



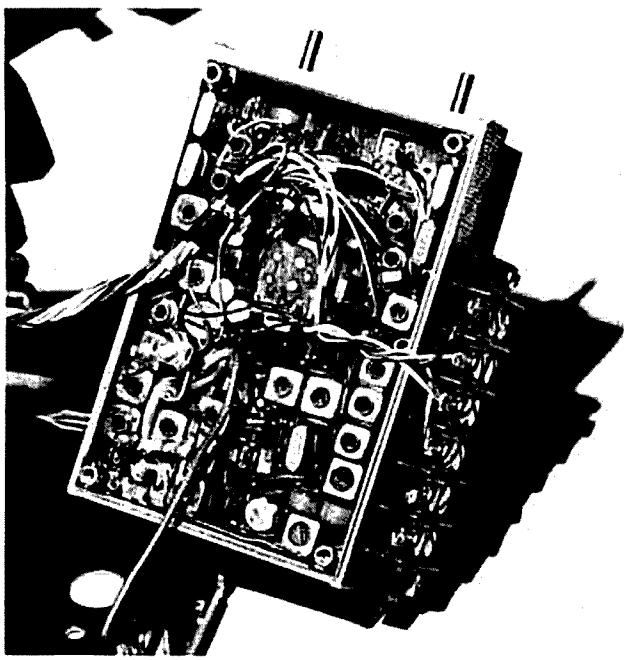
Mounting dimensions for volume and squelch controls.

lead dress. All the above leads will either be extended or replaced. Use Teflon-covered wire for extensions if possible. Unless you're experienced in working on printed-circuit boards, it is recommended that you extend the wires rather than try to run new leads from the board. Make a chart showing the relationship between original wire colors (on the board) and those you use to make the extension. Use a 3/8-inch-long piece of 3/64-inch-diameter heat-shrinkable tubing over each splice. Now continue as follows:

1. Clip black, green and brown leads 1/2 inch from volume control and extend each lead 9 inches. These leads will be

later connected to a volume control on the speaker.

2. Unsolder the red leads from the on-off switch on the volume control. Unsolder and remove the red lead connected to the pin jack on the board. (This pin jack *will not* be used for the + battery connection.) Join and extend the two red leads removed from the switch to about 3



HT-220 chassis assembly (top view).

inches. (One red lead goes to the relay; the other goes through an rf choke to the circuit board.)

channel selector. The channel-selector switches are separate 12-position Grayhill switches which are used for selecting the transmit and receive channels. The two transmit and two receive channels on the 220 circuit board are connected to positions one and two as shown in the schematic diagram in fig. 1. It is not necessary to strap the selector switches as is required when one switch controls both transmit and receive channels. This is because you can select any channel to transmit and/or receive, independent of one another. When you have completed wiring the selector switches make a chart showing the crystal frequency of each transmit and receive channel position.

3. Install the crystals as follows: Y-1 (transmit) and Y-101 (receive) nearest the end of the board; Y-201 (transmit) and Y-301 (receive) next to Y-1 and Y-101.

4. Extend the brown, yellow and purple leads 7 inches. These leads will be connected later to the squelch control.

5. Extend the following leads 4 inches: black/red control lead from the relay and the black (ground) lead from board next to the relay.

Note that the transmit section runs vertically along one side of the board and is easily identified by the final amplifier power transistor. The receive section runs along the opposite side of the board.

crystal-deck circuit board

The crystal deck was designed with the following objectives in mind:

1. To be able to peak up the HT-220 circuits with the crystal deck in place.

2. Isolation of crystals and peaking coils from the transmitter rf power output transistor.

These were met by using the layout shown in the photo showing the finished crystal deck wired to the selector switches. Receive crystals are located in the two left-hand vertical crystal holders and in the two left-hand positions of the upper horizontal molded rubber crystal holders. The transmit crystals are located on the right-hand side of the board. Note that the lower right-hand corner of the board is not used (it is directly over the transmitter power output transistor).

assembly. The circuit board is a 2-1/4x3-1/2-inch piece of fiberglass reinforced epoxy circuit board stock. A parts list is presented in table 2. The trimmer coils for each crystal are press-fitted into the holes in the circuit board. Note that holes have been drilled into the crystal deck board directly above the coils on the HT-220 circuit board so that the entire coil set can be peaked without disassembling the crystal deck. Press-fit term-

inal lugs are located along the edges of molded rubber crystal holder. Short lengths of no. 18 solid copper wire are used as terminal lugs for the connections between the end of the trimmer coil and the lead wires to the band switches. Using pliers, pinch the soft copper wire on each

It is recommended that the wiring on the circuit board be completed before connecting the leads to the two selector switches.

The layout shown worked out well and met the objectives. If you eliminate the need to tune up the HT-220 circuit

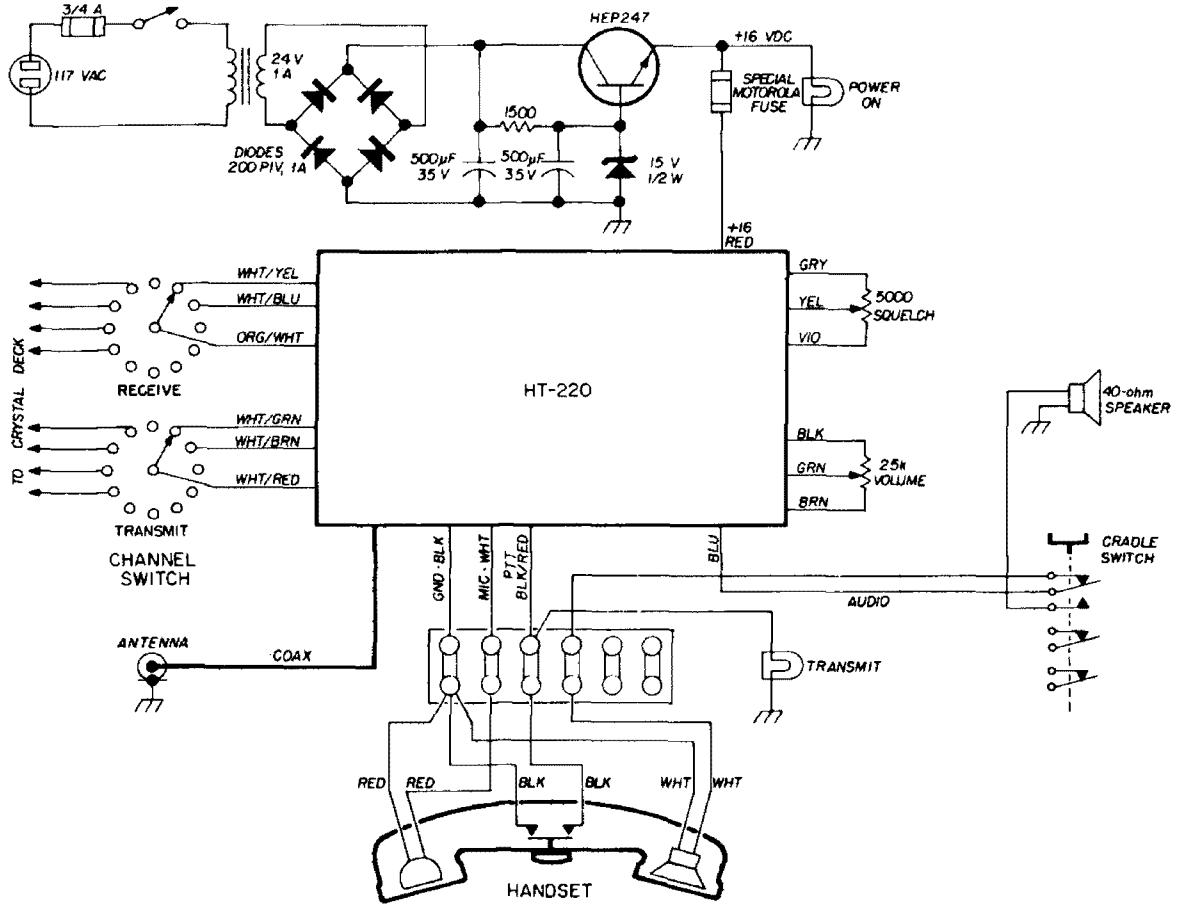


fig. 1. Power supply and interconnecting circuits for the HT-220 desk set base station.

side of the circuit board to lock it in place. Number 16 solid-copper wire is also used as the ground strap between the molded rubber crystal holders.

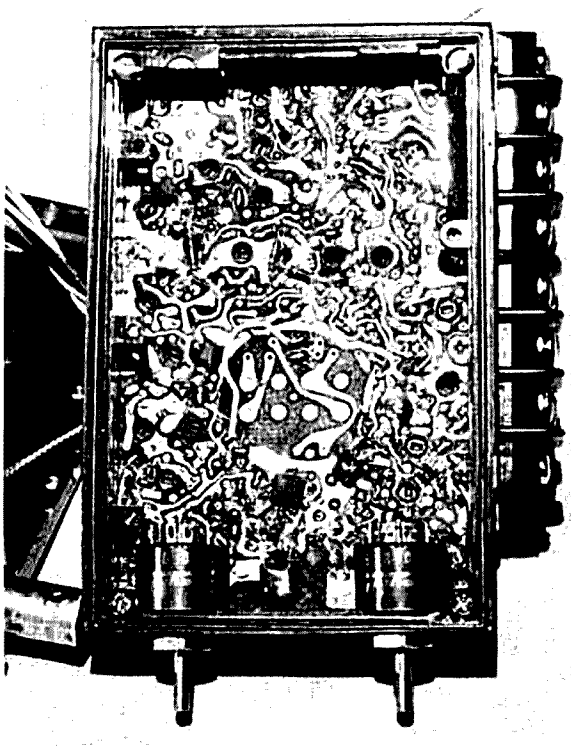
The 1/8-watt resistors in parallel with each crystal are mounted over the top of the molded rubber crystal holders between crystals. They will just fit. Don't try to use a resistor larger than 1/8 watt. The terminal at the very top of the circuit board between the two selector switches is connected to the discriminator circuit to accommodate a high-impedance meter, which can be used to zero in each receive crystal on the channel.

board without removing the crystal deck, considerably more leeway can be used in the layout.

One word of caution: with transmit and receive channels on separate switches, you must double check each time you change channels to be sure you have both the transmitter and receiver on the proper channels for the repeater you are working.

power supply

Fifteen volts at 500 mA are supplied by the power supply. Note that the transformer recommended (table 1) must



HT-220 circuit board (bottom view).

fit under the handset cradle switch. If you substitute another type of transformer, make sure it will fit into this space. All other components, except the voltage regulator transistor, are mounted on a perf board located under the speaker (see photos for details). A 15-volt zener controls the regulator transistor, which is mounted on a metal heat sink at the left-hand side. The regulated power transistor is insulated from the heat sink with a thin mica sheet. Use a transistor mounting kit and template to prepare the heat sink and to mount and insulate the transistor. Silicone grease should be used between the transistor and metal heat sink.

Note the use of component lead wires through the holes in the perf board for mounting and wiring the components. Mount the ac line switch and fuse holder. Ribs inside the plastic case must be cut off with a sharp wood chisel to accommodate the line switch and fuse holder. Next, connect the line cord. The lug on the end of the fuse holder may have to be partially cut off and insulated with tape to clear the HT-220 chassis.

Leave a service loop so the top plastic cover of the desk set can be removed while working on the set. Also, note that all leads except the antenna coax are in one cable so that the chassis can be easily removed for servicing.

miscellaneous mods

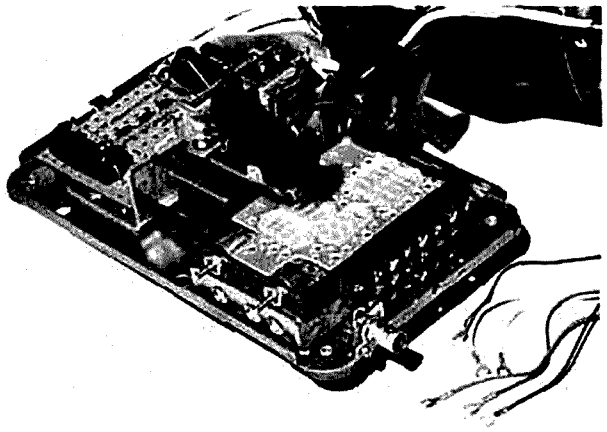
microphone. The handset carbon mike is removed and a Motorola miniature magnetic cartridge unit (table 1), made for the HT-220, is substituted. See photo for mounting details.

speaker. The handset is wired so that the speaker is enabled when the handset is in its cradle, and audio is switched to the earpiece when the handset is lifted from the cradle. The 40-45 ohm, four-inch-diameter speaker must be thin enough so it won't interfere with the cradle-switch level arm. Although the 4-ohm speaker supplied with the handset has a special cutout in the metal support arms, this speaker can't be used in this application because its voice coil impedance is too low for the HT-220 set, and its magnet is physically too large and will interfere with the power supply circuit board.

top cover assembly. To assemble the top cover on the desk set, remove the screw on the HT-220 chassis at the end nearest the right-hand side of the base, and

table 2. Crystal deck parts list.

quantity	item
1	3 x 4-inch fiberglass reinforced epoxy circuit board (Circuit-Stik Quik Circuits part no. 9354 or equivalent)
10	Motorola 24-83638H14 coils - (transmit)
10	Motorola 24-83638H13 coils - (receive)
5	Motorola 7-84274H01 molded rubber crystal holders
10	Motorola 6S185B78 1/8-watt 820-ohm resistors (receive)
10	Motorola 6S185B75 1/8-watt 470-ohm resistors (transmit)
21	press-fit PC board solder terminals
2	flat head Phillips screws
1 ft.	no. 16 bare copper wire for ground straps and coil terminals
20 pcs.	no. 26 gauge Teflon insulated hook up wire, 5-inches long



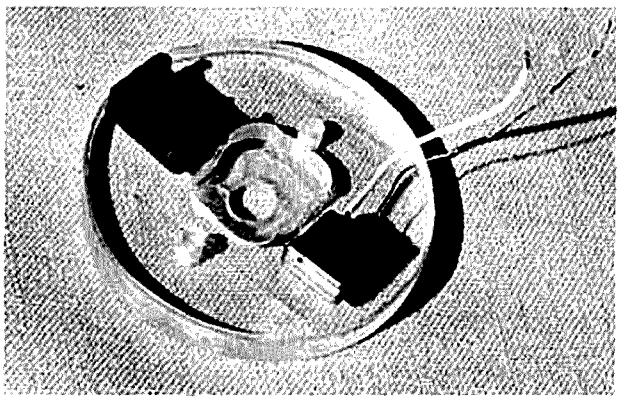
Crystal deck circuit board, BNC coax antenna connector, and terminal strip for handset.

loosen the rear screws so that the HT-220 chassis can be tilted upward to get the channel switch shafts through the holes in the right-hand side of the plastic telephone-set housing. The cradle levers must be manipulated through their slots, and the front hook on the bottom base must engage the slot at the front of the cover. It may be necessary to wedge the hook in place with a screwdriver.

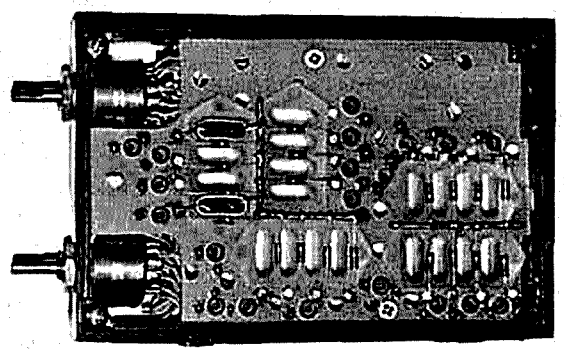
Check the cradle switch to ensure it works freely after the screw at the rear of the desk set housing has been tightened. Note that the on-off switch bottom mounting lug should be on the inside of the lip on the base. Loosen this screw while assembling the set and retighten. Install the channel switch knobs.

rf section. Small padding capacitors are sometimes needed in the driver and final-amplifier circuits. Add these capacitors, one at a time, as you go through the tuneup procedure and determine if they help to increase power output. Monitor the dc input to the HT-220 board during tune up, and do *not* keep the power on for more than a few seconds if the current exceeds 425 mA, since there is danger of damaging the driver transistor. Adjust final-amplifier tuning to reduce current to 400-425 mA.

When feeding a remote power amplifier, limit the length of small coax cable used at the output of the desk set to five feet or less. If you must run a longer cable to the remote power amplifier, use larger cable; preferably RG-8A/U. The short length of small-diameter cable at the desk set is only meant to give flexibility in moving the radio around on the desk.



HT-220 miniature mike cartridge, molded rubber holder, and plastic adapter plate.



Assembled crystal circuit board connected to selector swtiches.

final remarks

Schematics and tune-up instructions should be obtained when you buy your board (available at nominal cost). The matching filter between the coax feed line and antenna is not necessary; the match is satisfactory without it.

Name plates for the channel switches, volume control, and squelch control help dress up the desk set. Your call letters could be substituted for the Motorola name escutcheon on the speaker grill.

This is a very nice rig. While it's a narrowband set (5 kHz), it accepts 8 - 10 kHz deviation without much distortion and sounds fine on most repeater operations.

ham radio

television DX

Chasing broadcast television DX is a step above short-wave listening, and provides some interesting data on vhf propagation modes

Morrie S. Goldman, WA9RAQ, 5815 North Christiana Avenue, Chicago, Illinois 60659

There is a usual pattern leading up to the time when a person becomes a ham. Many of us started out by listening to a distant station on an a-m broadcast radio, only to graduate up to shortwave listening. And finally, often with the aid of an amateur aquaintance, we worked for, and earned, an amateur radio license.

My own interests followed a somewhat different route. I, too, went through the listening stage, but I was rather intrigued with *seeing* DX as well as hearing it. What I am referring to of course is *TV-DX* — receiving normal television stations, during abnormal conditions, over relatively great distances.

Television DX is made possible by the same propagation vehicles as DX on the vhf and uhf amateur bands. Band openings are just as exciting, and with the added dimension of video, far more dramatic. Somehow, *watching* 1000-mile distant stations float in with sparkling clarity or even heavy snow has an attraction all its own. Suddenly, you're *seeing* heterodynes, meteor bursts and auroral flutter. Whether the signals are originating from a five-megawatt station or a low power relay, the excitement is still there.

As you will see, television DXing can also be of practical advantage to the amateur. Whether it is used as a research tool or simply as an early warning device for vhf and uhf openings, television DX should be more than a novelty to the serious vhf enthusiast. This aspect of TV DX will be covered later in this article.

propagation

Without exception, the same types of propagation that affect the vhf ham bands affect the TV broadcast channels. Since much has already been written in handbooks and magazines about vhf and uhf propagation, and because of space limitations, I have chosen to include only brief details here. Several references to more detailed studies of vhf propagation are included at the end of this article. It is, however, worth pointing out that you should avoid older texts on the subject. Much of what we know about vhf and uhf propagation has only been detailed in the past 15 years or so. Most earlier texts are full of false assumptions. The articles cited will provide the most up-to-date information.

Tropospheric bending Tropo, which is sometimes known as extended ground wave, is a weather related phenomenon affecting both the vhf and uhf television

channels. Several variations of this propagation mode exist, all requiring the influence of a high-pressure area. Basically, when a temperature inversion takes place in the troposphere, a strata developes which forces vhf (and higher) signals to follow the curvature of the earth. This continues for the length of stratified layer, anywhere from 60 to 1000 miles.

Sometimes two layers form close enough to each other to simulate a wave guide, *ducting* signals for 350 to 1000 or more miles. *Tropospheric ducts* often channel signals over all points between the transmitter and receiver, much like skip. Since the duct is a waveguide, it may affect a relatively small frequency range (such as part of the uhf band), or it may be large enough to affect the entire TV spectrum. Ducts are most common along the edge of a weather front.

The warmer months and especially the autumn are best for tropo, although it can occur during any season. Best times are usually around sunrise and sunset when rapid heating and cooling in the troposphere takes place. Tropo is characterized by steady signals and slow fading.

Sporadic-E skip (Es) Also known as short skip in amateur circles, Es is most common in the late spring and early summer months, with a secondary lesser peak occuring from mid-December to early January. Best times are most often from mid morning to early afternoon and from early evening to about 2230, local time. Sporadic-E occurs when vhf signals hit sporadic patches or *clouds* of ionization in the E layer of the ionosphere and are refracted to a distant point. Typical distances for a single hop range from approximately 450 to 1400 miles.

Double and even triple hops are possible and have been reported, but are not very common. Es begins low in frequency and as the density of the patch increases, so does the muf. Es commonly affects all of the low band (channels 2-6) and occasionally extends to channels 7 and 8 in the high band. Es is the most unpredictable and misunderstod of all the types of ionospheric propagation.

Sporadic-E propagation is characterized by strong signals and deep fading. Reception can last for minutes, hours and in some rare cases, days.

F2 layer skip (F2) During years of peak sunspot activity the F2 layer of the ionosphere can provide a daily reflective barrier to vhf signals. Skip distances begin at about 1700 miles. With multiple hops, the ultimate range is unlimited. During the past sunspot peak, the muf seldom reached our channel 2, but did bring in 41-MHz TV audio signals from France and England. In 1957, George Palmer of Williamstown, Victoria, Australia, set the world's TV-DX record by receiving BBC-TV audio by long path F2 over a distance of 14,500 miles. The video was received on another occasion during the same year by short path F2 over a distance of 10,400 miles!

Auroral Scatter (As) The aurora borealis or northern lights does more than wipe out low-frequency DX – it causes DX to occur at vhf. Though once thought to be relatively well understod, auroral scatter is of increasing interest to researchers. It can propagate signals anywhere from several hundred to several thousand miles. Aurorally propagated signals are characterized by rapid flutter. In fact, this flutter is usually so intense that it garbles television and audio signals beyond recognition. Only during a small percentage of the magnetic disturbances are signals stable enough to be useful.

Peak signals for auroral scatter DX appear most often toward magnetic north. Reception and visual auroral sightings are far more common in the northern states and Canada than in southern regions.

Auroral peaks seem very closely tied to sunspot peaks. According to ionospheric physicist Gordon P. Nelson, "magnetic and solar activity may lead or lag each other in alternating cycles."* The

**Auroral Disturbances:* A talk presented by Gordon P. Nelson at the 1972 convention of the Association of North American Radio Clubs in Boston, Massachusetts on July 15, 1972.

magnetic peak lagged the solar peak in cycle 20, so we would expect the reverse to be true in cycle 21.

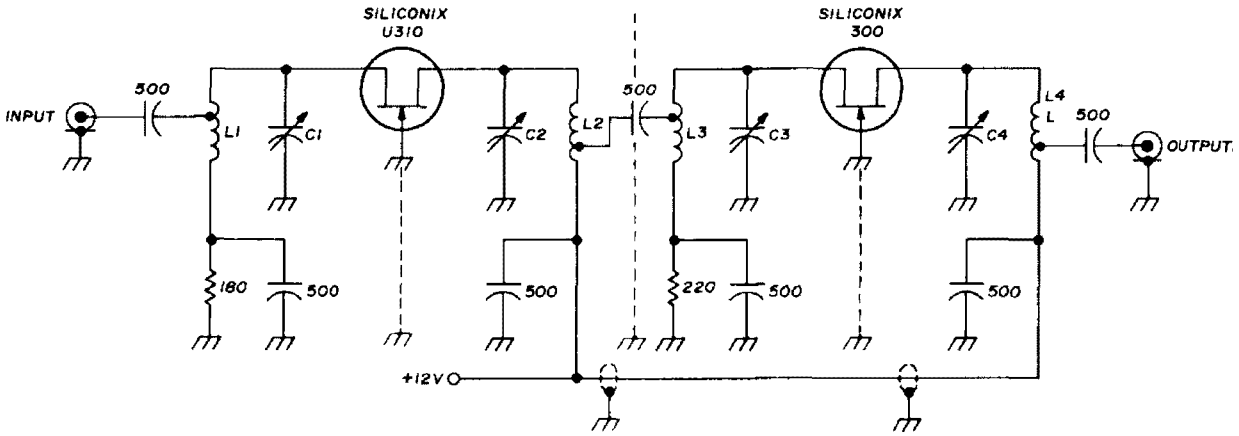
Meteor scatter (MS) When a meteor is pulled toward the earth and begins to burn from friction, it leaves an ionized trail behind. This short-lived ionization occurs in the E layer of the ionosphere and creates momentary bursts of E skip. Meteor scatter bursts can last anywhere from a fraction of a second to over a minute and produce DX from about 1100 miles to as close as 180 miles. Most common are signals in the 500- to 900-mile range.

Best results can be expected during meteor showers, when large swarms of meteors enter the earth's atmosphere. (A list of showers can be found in *The Radio Amateur's VHF Manual,* published by the ARRL.) However, there are always sporadic meteors being pulled into the earth's atmosphere.

For TV-DX purposes, the best time for meteor scatter is from about 0400 to 0800 local time when many TV stations are displaying test patterns and ID slides. From midnight until noon, local time, we are at "the front of the earth," catching the greatest number of meteors with the greatest force. There are some daylight showers, but on an average, the greatest activity is between midnight and noon.

With average equipment, a great deal of meteor scatter can be observed on the low band TV channels. For results at channel 7 or higher, stacked, cut-to-channel antennas and a low-noise preamp are required.

Lightning scatter (Ls) Lightning scatter is a very interesting and recently discovered mode of propagation. In fact, the discovery is so recent that almost no research has been conducted into its mechanics. This much *is* known — when very intense thunderstorms are between two points, perhaps from 200 to 500 miles



	low-band		high-band
C1,C2 C3,C4	3-32 pF variable (E.R. Johnson 167-130-1)	C1,C2 C3,C4	1.5-5 pF (E.F. Johnson 167-102-1)
L1	5 turns no. 16, 3/4" diameter, tap down from hot end for best performance	L1	3 turns no. 16, 3/4" diameter, tap down from hot end for best performance
L2	6 turns no. 16, 3/4" diameter, tapped 3/8-turn from cold end	L2	3 turns no. 16, 3/4" diameter, tapped 3/8-turn from cold end
L3	5 turns no. 16, 3/4" diameter, center tapped	L3	3 turns no. 16, 3/4" diameter, center tapped
L4	6 turns no. 16, 3/4" diameter, tapped 1/8-turn from cold end	L4	3 turns no. 16, 3/4" diameter, tapped 1/8 turn from cold end

fig. 1. A tunable fet preamplifier for use on vhf TV. For higher performance on the high band (channels 7 through 13), additional stages identical to the second stage may be added.

apart, they can reflect uhf signals between those points. Exactly how the signal is propagated, or why lower frequencies are not affected, has not yet been determined. It is believed that the first recorded observation of lightning scatter was made in June, 1968, by a TV-DXer. In September of that same year W5RCI and WØDRL made the first 432-MHz contact via lightning-scatter propagation.*

Reception during lightning-scatter propagation appears very much like that of meteor scatter. Signals pop in and out with varying strengths. I have even seen signals reach near snow-free levels for several seconds. If any readers have further data on lightning-scatter reception, I would very much like to hear from them.

equipment

Just as in amateur radio, how serious you are about the hobby will determine what caliber of equipment you use. At the bottom of the scale are the occasional TV-DXers, many of whom only choose to use indoor rabbit ears! This is fine for a touch of strong summer sporadic-E, but not much more.

On the average, most TV-DXers are using a quality consumer TV set and a broadband Yagi or log-periodic vee antenna. The antennas are usually fringe-area models.

For uhf reception a separate antenna, of any of a number of types, is mounted on the same mast. The array is rotatable and usually mounted anywhere from 30- to 50-feet above ground. Where overloading is not a consideration, mast-mounted vhf and uhf preamps may be used.

The more serious TV-DXer uses a separate Yagi or log-periodic Yagi[1] for the low and high bands. He often uses stacked antennas for increased gain, capture area and directivity. Single-channel antennas are also used for maximum results on a given channel. At vhf, low-loss RG-59/U cable (such as Belden Duo-Foil or International Interfoil 750) is used throughout. For the serious DXer, consumer-type preamps are now taboo because of their susceptibility to overload, the resulting cross-modulation, and

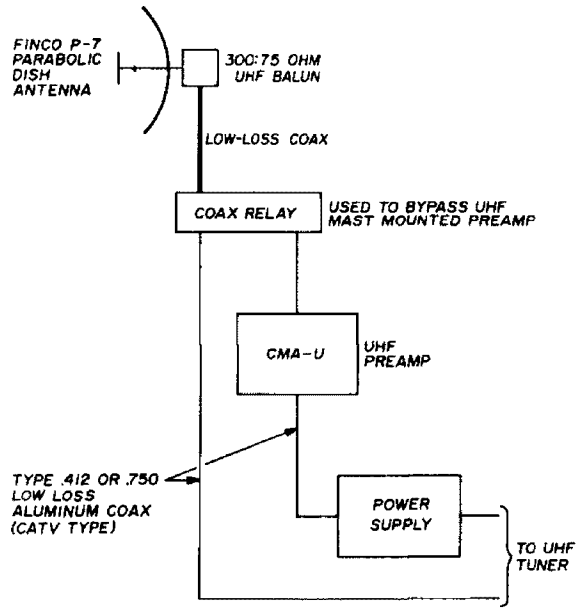


fig. 2. A typical high performance installation for DX reception of uhf TV stations.

poor noise figures. For the vhf TV channels the serious DXer uses a home built tunable, multi-stage fet preamp similar to the one shown in fig. 1. This preamplifier improves adjacent channel rejection and limits video bandpass as well as providing improved noise figure and high gain. Even though there are many tuning knobs, once the dials are calibrated, tuneup is quite fast.

On the uhf TV channels (470 to 890 MHz) a seven-foot parabolic dish antenna is a must. In the reasonable price range the Finco P-7 is about the best available. Once again, consumer type preamps are not used. Their noise figure is often no better than the uhf tuner or converter.

However, there are several CATV-type, broadband uhf preamps that *will* help. Here, the best buy is the Blonder-Tongue CMA-U. Unfortunately, in strong signal areas, overload and cross-modulation can

*The first recorded observation of lightning scatter propagation was described in the August, 1968, issue of the *VHF-UHF Digest.* The two-way 432-MHz lightning-scatter contact is discussed in "The World Above 50 MHz," *QST,* November, 1968, page 89.

occur on these units as well. To combat this, I installed a remote 75-ohm coaxial relay, such as the Dow Key 77-2632, to switch the preamp in and out, as needed (see fig. 2).

The uhf antenna and, hopefully, the high-band antennas are mounted a minimum of 50-feet above ground. A minimum of 25-feet should do for the low band antenna. To combat other assorted local problems, the serious TV-DXer uses bandpass filters, tunable traps, attenuators, combline interdigital preamplifiers[2] and other devices.

The receiver is still generally a high

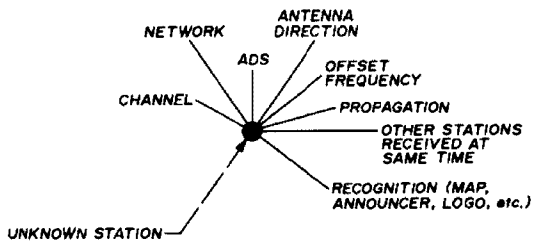


fig. 3. The identity of an unknown TV station can often be determined if there are enough pieces of information (see text).

quality consumer type, but a number of TV-DXers (myself included) have gone to more elaborate CATV-type receivers. These sets boast such features as manual rf gain controls, variable agc positions, video level controls, four i-f stages and other refinements. They do however, require the addition of a video monitor for visual display and an audio amplifier for sufficient audio to drive a speaker.

identifying unknown signals

Often, the TV-DXer finds it necessary to play Sherlock Holmes in an attempt to identify an unidentified signal. Searching for clues, he tries to fit the pieces of the puzzle together. Fig. 3 illustrates some of the required considerations. If enough clues point to one station being the prime suspect, the TV-DXer makes his decision. However, as any good detective or DXer knows, you must not jump to conclusions with insufficient evidence!

One consideration is called *offset.* This refers to whether a station is assigned to be exactly on channel or is offset, 10-kHz high or 10-kHz low of center. The FCC assigns offset frequencies to minimize co-channel interference. As a side benefit of this, the offset patterns produced when two stations beat against each other can reveal valuable information about one of the stations.

For example, if you are receiving a known station (call it A) on channel 12 and see another station (X) beating against it, you can determine the offset frequency of station X by the offset pattern observed. From a station list, you can determine that station A is assigned to operate on -12.* If the heterodyne of co-channel station X is producing many fine horizontal lines, station X is 20 kHz away (+12). About 10 ot 15 larger bars indicates a 10-kHz separation (12 exactly); 5 or 6 still thicker bars indicates zero offset (−12). Thus, if the reference station (A) has a plus or minus offset, you could see any of three possible patterns from an interfering station. However, if station A is exactly on channel, you would see only two patterns (zero and 10 kHz).

Offset lists are not 100% accurate partially because of allocation changes, typographical errors and carrier drift (beyond FCC allowances). Most of the time however, offsets can greatly help in adding another piece to the puzzle.

Time zones are not listed in fig. 3 because many stations delay their broadcasts. This could give you a false impression that a station is farther, or closer (depending on whether you're located in the East or West) than it actually is.

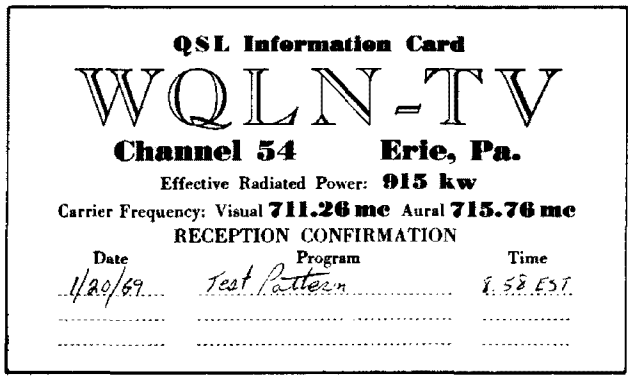
The *VHF-UHF Digest,* described later, features a column devoted to helping TV-DXers decide just what they received.

**Broadcasting Yearbook,* available from *Broadcasting Magazine,* 1735 De Sales Street, NW, Washington, DC 20036, lists all U.S. Television, fm and a-m stations (including tv offset frequency information). The *WTFDA TV Station Guide,* priced under $5.00, available from WTFDA, Box 163H, Deerfield, Illinois 60015, provides complete listing of U.S. tv stations by channel.

The DXer sends a detailed report of his reception to the column editor, who in turn scans *TV Guides* from around the country. If the editor comes up with a blank too, the report is printed in the hopes that another DXer will recognize the logging and its origin.

confirmations

Amateurs and radio broadcast stations are not alone in receiving QSL requests. Many TV-DXers send their reception data to TV stations and request a written verification. Most stations do verify, and now many TV stations even have their own printed QSL cards. The chief engineer is usually the "verie signer" and he's often a fellow ham. Most veries (verifications) are still typewritten letters and often carry some rather amusing reading. Such lines as, "We hope you will continue watching our station," "This will verify receipt of your report," and, "We wish you were one of our regular viewers," pop up frequently. There's even the occasional, "No, you would not have been able to receive our station, because our new antenna transmits no sky waves."


QSL Information Card

WQLN-TV

Channel 54 Erie, Pa.

Effective Radiated Power: 915 kw

Carrier Frequency: Visual 711.26 mc Aural 715.76 mc

RECEPTION CONFIRMATION

Date	Program	Time
11/20/69	Test Pattern	1.58 EST

fig. 4. QSL card is typical of many used be other TV stations. This card is signed on the reverse side by Assistant Chief Engineer, WA3DLS.

photographing TV-DX

Perhaps this section should be subtitled, "how to get an instant QSL." What better souvenir of a DX logging could you have? TV-DXers have long been known to photograph test patterns and ID slides of DX stations as instant proof of reception. The photos are also great for making believers out of skeptics.

It's a simple matter to photograph TV-DX once you know the basics. If your camera has a leaf-type shutter, a setting of 1/30 second at around f/5.6, using Kodak Tri-X film, usually works out well. For cameras using a focal-plane shutter, try 1/8 second at f/8.

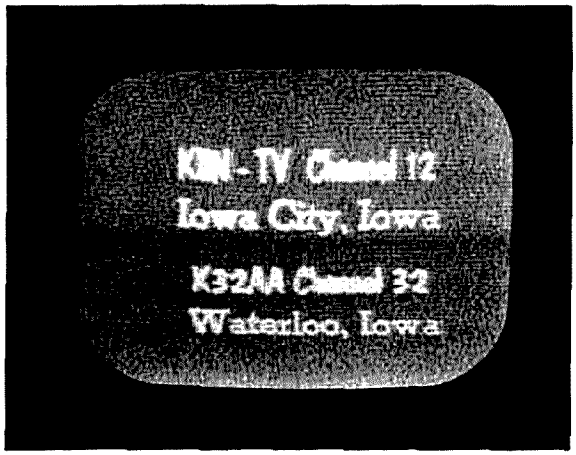
Sight and focus the camera so that the entire screen is clearly in view. *Never* use flash, the light bounces off the face of the picture tube, giving you an obliterated photo. For the same reason, reduce the amount of external light striking the screen as much as possible. To prevent camera motion, the use of a tripod is highly recommended.

A highly detailed 8-page booklet entitled, "Photographing Television Images," is available free of charge from the Eastman Kodak Company (Consumer Markets Division, Rochester, New York 14650). Just request Kodak Customer Service Pamphlet AC-10. This booklet describes photographing black-and-white and color TV images using a variety of cameras including movie cameras.

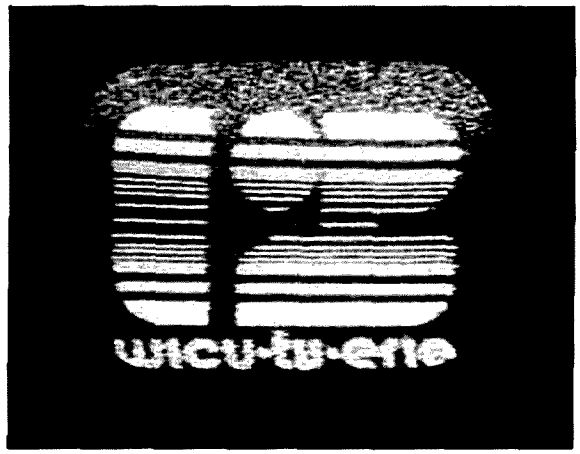
One trick you won't find listed in the Kodak pamphlet is how to turn snowy pictures into clear ones. By using a time exposure of 1/2 to several seconds, you can clear the snow out of a weak picture. Remember to adjust your exposures accordingly. If you have a lightmeter it can be used effectively for TV images as well. As with all forms of photography, practice will improve your results.

vhf amateur applications

Unless you're already too intrigued with TV-DXing itself, you've probably started to think about its amateur applications. The most obvious of these is as a propagation monitor. You may notice the effects of a vhf opening on your TV set before you even turn on your vhf rig. And secondly, once operating, TV-DX observations can aid in determining the geographic coverage of an opening. When used in conjunction with an fm broadcast



Reception is of KIIN-TV, channel 12, Iowa City by tropo. 210-miles distant.



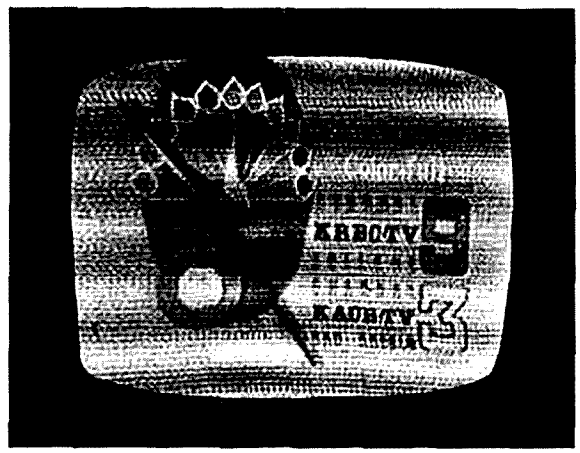
WICU-TV, Erie, Pennsylvania. 400-mile reception by tropo.

radio, these observations can also help monitor the muf. One practical application of this technique is to determine

and a high-gain parasitic array. A TV set will also work out quite well if it is used with a low-noise channel-7 mast-mounted



KTVC, channel 6, Ensign, Kansas. Received by sporadic-E at 725 miles.

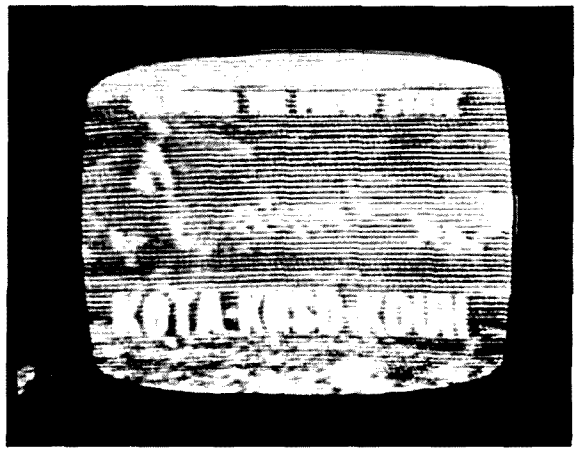


Another sporadic-E catch KACB-TV, San Angelo, Texas. Two offset patterns are seen in this photo. The large black bars indicate zero offset, and the fine lines, a 20-kHz pattern. 1010-miles distant.

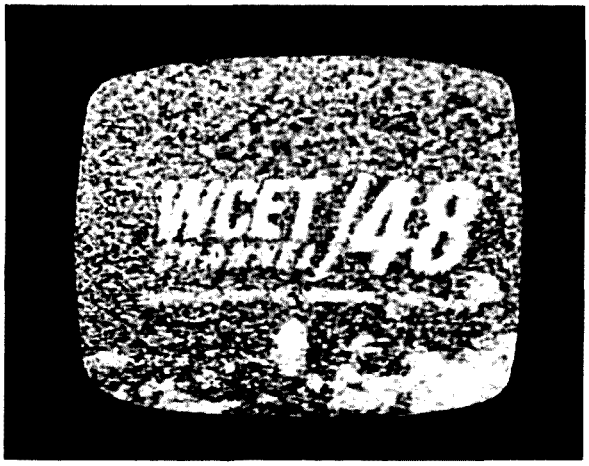
whether and when 2-meter sporadic-E is likely. The same is also true of auroral DX.

Several avid 2-meter meteor-scatter enthusiasts have realized that they can stay one step ahead by monitoring the activity on channel 7. As a meteor enters the earth's atmosphere, its ionized trail starts small and rapidly grows longer. Therefore, the trail more readily reflects higher than lower frequencies. In short, a meteor burst is very often heard on TV channel 7 just before it's heard on 2 meters.

For this application, there are two possible easily duplicated monitoring systems, a low-noise crystal-controlled converter tuned to about 179.25 MHz



Reception is of KOTA-TV, channel 3 in Rapid City, South Dakota, by sporadic-E. Note the fine lines indicating 20-kHz offset co-channel interference. 785-miles distant.



WCET, channel 48, Cincinnati, Ohio. A 255-mile tropo path.

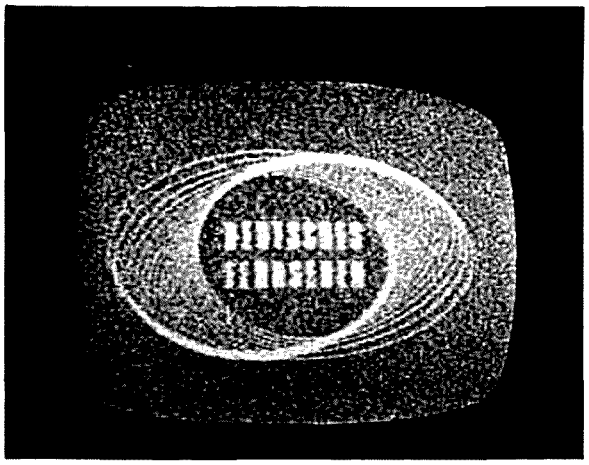


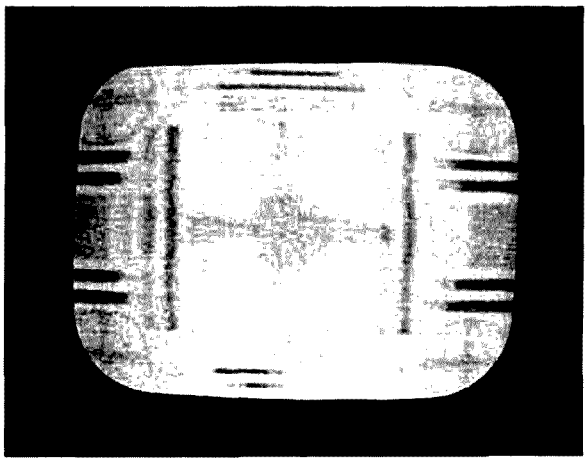
Sporadic-E reception of CBHT-TV, Halifax, Nova Scotia. 1189-miles distant.

preamp.* In fact, with a TV receiver, you stand a good chance of identifying the station bursting in. Either way, the single-channel mast-mounted preamp is a great asset.

If you're looking for a vhf field-day site, a battery-powered TV set can be your best friend. It will aid in determining how good a location is for vhf reception. Monitor some of your fringe stations at each trial site. You may find that results can be greatly enhanced by moving only a small distance.

Whenever I start thinking about the correlation between TV-DX and 6-meter propagation, I can't help remembering a rather intense sporadic-E opening that occured several years ago. While TV DXing on channel 3, I began to hear some TVI over the TV-DX station from an





TV-DX is also popular in Europe. These photos were taken by a DXer in Holland. Top: A.R.D., West Germany, by tropo. Bottom: Belgium by tropo on channel E2.



KS2XBR, channel 38, Chicago, is one of the last experimental TV stations in the United States. This station is operated by Zenith Radio for testing pay TV devices. Truely a rare catch.

*The CADCO IPA-HB28/7 preamp is probably the unit best suited to this purpose. The noise figure of this prearnp is under 1.5 dB and gain is 28 dB. Information on this preamp is available from CADCO, Inc., Box 18904M, Oklahoma City, Oklahoma 73118.

amateur in California. As if hearing double-hop TVI wasn't enough of a shock, I discovered that his signal was stronger on the TV than on 6 meters! I dropped this fellow amateur (who shall forever remain unnamed) a note, and received a reply written on the back of a QSL card. The reply stated that I was about the 50th person to complain about



Author's home station is used for TV-DX as well as for amateur communications. Note CATV-type TV monitor in upper right.

his TVI – the other 49 were his neighbors. Anyway, I was the one to finally convince him that his neighbor's TV sets weren't at fault!

TV-DX club

In the United States there is one organization devoted to TV, fm and vhf utility DX – The Worldwide TV-FM DX Association. Monthly, WTFDA publishes the *VHF-UHF Digest* which features columns devoted to member's DX reports, FCC and CRTC news, theory, construction, statistics (DX records), QSLs and others. Feature articles frequent the winter issues. You might say that WTFDA serves the same function for non-amateur VHFers as the ARRL serves for amateur radio operators. WTFDA even holds annual conventions. This non-profit group strives at promoting vhf and uhf research and technical advancement and prides itself on its very devoted members.*

Mel Wilson, W2BOC, is a WTFDA member and credited the organization with providing detailed sporadic-E reports in the *VHF-UHF Digest.* These reports provided Mel with hard data to add to his 30+ years of propagation research.

Also well known in vhf circles for his propagation studies is Pat Dyer, WA5IYX. Pat is WTFDA's "VHF Utility DX" editor. His column covers just about every form of non-broadcast, non-amateur DX above 30 MHz. By reviewing reports in Pat's column, it is easy to follow the rise and fall of the F2 muf.

Whether or not your main interest in amateur radio lies in vhf, during the coming spring and summer months I hope that you'll keep one eye on those *empty* TV channels. Who knows, maybe you'll lock onto some TV-DX and get hooked, the way I did!

*A one-year membership in the WTFDA is $7.00 and includes 12 monthly issues of the *VHF-UHF Digest* (for first-class mail delivery of the *Digest* the annual dues are $9.00). Sample copies of the *VHF-UHF* Digest are available for 50 cents each, postpaid. Rates for subscribers outside the United States are available upon request. Write to the WTFDA, Box 163-H, Deerfield, Illinois 60015.

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ham radio

how to select batteries for portable equipment

A discussion of the performance and size and cost characteristics of the many types of batteries currently available

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Inexpensive transistors and multifunction integrated circuits have made possible a whole new generation of portable and low-power equipment for the radio amateur. They have also relieved many construction headaches for the ham who builds his own equipment. One of those headaches used to be power supply construction, but now batteries are the most popular power source because of their simplicity, low cost and portability. However, whether you build or buy your equipment, one big problem remains — selecting the batteries. The selection process may be rather involved, particularly for the builder, who has more options open to him.

battery characteristics

The first step in selecting a battery is to determine in detail the power requirements which the battery system must meet. You also should have an idea of which requirements are most important. This can be illustrated by example: Suppose you are building a portable transceiver which will use a 9-volt dc power supply. It will draw about 100 mA while transmitting and 30 mA while receiving. A list of battery parameters, in order of their importance, might look like this:

Battery voltage. This is first on the list for obvious reasons. There isn't much point in examining all the available 24-volt batteries when you only need 9 volts. But even this important parameter isn't as straightforward as a first glance might indicate. Since the output voltage of a

battery declines with operating time, you must specify the minimum and maximum supply voltages the transceiver can tolerate. Let's say that 7.2 to 10.0 volts is acceptable.

Power capability. This is almost as important as battery voltage. If the battery simply will not deliver 100 mA, it cannot be considered.

Cost. Although we all want to stretch a dollar as far as we can, you must weigh the expected battery life against its initial cost.

Size/weight. The importance of a battery's bulk will vary with the type of operating you do and the size of the radio itself. In the case of the example, you may want to trade off increased size and weight for decreased cost, if necessary. You should, however, try to stay under six pounds and 100 cubic inches.

Battery life. The economics of this parameter are covered above, but the length of time between replacements or recharges is also a matter of convenience. You may want to operate for most of a weekend without worrying about the battery giving out.

Availability. If you choose a battery type which must be replaced frequently, you must be sure that replacements are easily obtained.

Others. Some other battery parameters which are not important for the example are high and low temperature characteristics, ease of storage and long shelf life.

examining available alternatives

Let's see what is available to choose from. Since the *carbon-zinc* battery or Leclanché cell is the most widely used type of battery, and usually the least expensive, look there first. Many of the 9-volt carbon-zinc batteries are not capable of supplying 40 to 50 mA, average, for any appreciable length of time. (Estimates of battery life are based on 50 mA average current, which would correspond to transmitting for about 30% of the operating time.) One battery that will is the Eveready 716 or the equivalent Burgess 4F6H – it should supply the transceiver for more than 300 hours. However, this battery costs over eight dollars, weighs 8½ pounds, and occupies over 200 cubic inches of space.

Another approach would be to use six common D-size flashlight cells in series. This could be the cheapest approach, but you would get less than 80 hours of battery life and battery volume would go up to about 340 cubic inches.

While I am discussing carbon-zinc batteries, I should mention the possibility of recharging them. Although such recharging can be done, it is probably not practical, and can be dangerous. Excessive gassing may result from too high charging current and the cell could explode. The following is an excerpt from National Bureau of Standards Letter Circular LC 965:

"From time to time attention has been turned to the problem of recharging dry cells. Although the dry cell is nominally considered a primary battery it may be recharged for a limited number of cycles under certain conditions. Briefly these are:

1. The operating voltage on discharge should not be below 1.0 volt per cell when battery is removed from service for charging.

2. The battery should be placed on charge very soon after removal from service.

3. The ampere-hours of recharge should be 120 - 180% of the discharge.

4. Charging rate should be low enough to distribute recharge over 12 - 16 hours.

5. Cells must be put into service soon after charging as the recharged cells have poor shelf life.

Recharging of dry cells may be economically feasible only when quantities of dry cells are used under controlled conditions with a system of exchange of used cells for new ones already in practice, and with equipment available to

provide direct current for charging. Such a system would not be practical for home use."

mercury batteries

It looks as though carbon-zinc batteries could satisfactorily fulfill the transceiver power requirements, but let's see what else is available. What about *mercury* (mercuric oxide) batteries? They have the advantages of a very stable cell potential and high power capability for a smaller size. They are also expensive.

Since a single 9-volt battery of sufficient power capacity is not readily available, let's look at six D-size cells (such as the Eveready E42) in series. Here you could approach 300 hours of operation, but at a cost of about thirty dollars and with a volume exceeding 300 cubic inches. Although it is possible to recharge mercury cells, decreasing the overall operating cost, manufacturers discourage the practice because of the danger of cell explosions. Apparently, mercury batteries will not fulfill the requirement.

alkaline batteries

The *alkaline* (alkaline-manganese) battery is becoming increasingly popular both because of its higher service capacity and because many types are rechargeable. Although the alkaline-manganese cell has the same 1.5-volt open-circuit cell voltage as the carbon-zinc cell, it discharges at lower voltages. The carbon-zinc cell provides most of its available energy above 1.25 volts and will be just about exhausted at 1.0 volt. However, the alkaline-manganese cell yields most of its energy below 1.25 volts and still has quite a reserve below 1.0 volt.

This means that if you use six alkaline-manganese cells for the 9-volt transceiver power supply, you should expect a significant percentage of operation to be at 6 volts or less. Such a supply voltage drop would not be acceptable for the

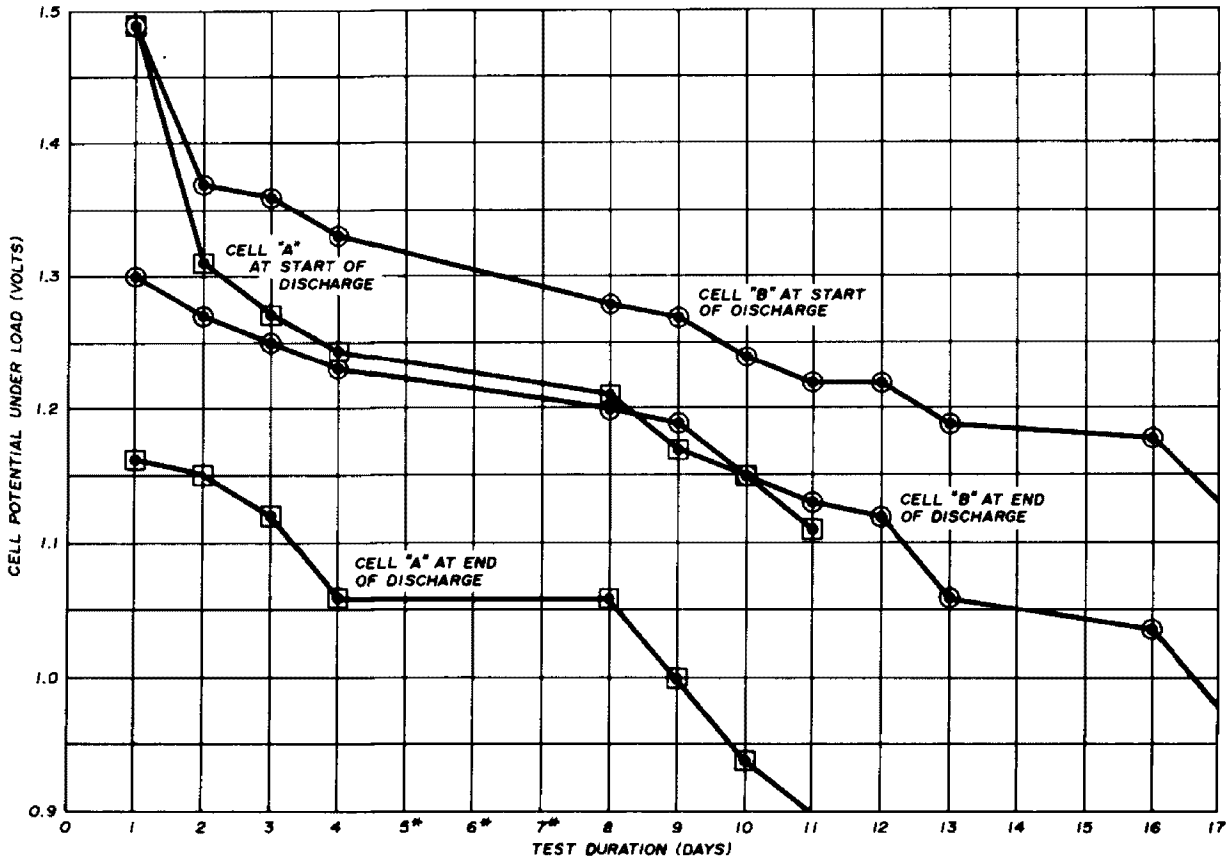


fig. 1. Comparison of two 1.5-volt carbon-zinc cells, discharged 3½ hours per day into a 10-ohm load. Cell A is a low-cost unit while Cell B is a so-called super cell, manufactured by a well-known company. On days marked with an asterisk the cells were not discharged.

transceiver in the example. However, if the system is to be recharged, discharge is usually terminated at about 1.1 volts. Thus, a rechargeable alkaline battery system would provide acceptable operating voltages.

I found no readily available 9-volt alkaline batteries, but rechargeable 1.5-volt D-cells are easily obtained. Six of these cells should cost about ten dollars and should last 30 - 40 hours between recharges. If battery life is 50 recharge cycles, then the initial investment should be good for nearly 2000 hours. The recharge operation can be simplified by purchasing one of the special chargers which are generally available for less than ten dollars. Size and weight are about the same as carbon-zinc D-cells. Rechargeable alkaline batteries are definitely among the possibilities for powering the transceiver.

table 1. Performance and cost of batteries used for a 9-volt power supply.

type	size (cubic inches)	weight (lb.)	estimated life (hours) per charge	estimated life (hours) total	cost initial	cost per 100 hours
9 volt carbon-zinc (Eveready 716, Burgess 4F6H)	204	8½	310	310	$ 8.95	$ 2.90
Six 1.5 volt carbon-zinc D cells	345	1½	280	280	$ 1.50	$ 1.90
Six 1.35 volt mercury D cells	345	3	280	280	$30.00	$11.00
Six 1.5 volt alkaline rechargeable D cells	345	2	40	2000*	$10.00	$ .50
10-volt nickel-cadmium (Eveready 1007)	63	4	80	40,000†	$40.00	$ .10

*Based on an estimated 50 cycle life
†Based on an estimated 500 cycle life

nickel-cadmium batteries

When you discuss rechargeable batteries, everybody thinks of *nickel-cadmium* batteries. This popular battery has led to the development of a tremendous variety of cordless, rechargeable devices because they combine the ease of portability with the freedom from having to buy fresh batteries everytime you turn around. As might be expected, all these good things do not come cheap. Nickel-cadmium D-size cells for use in the transceiver would cost over six dollars each, and would be just as bulky as the other types.

Another alternative would be to use a 10-volt nickel-cadmium battery such as the Eveready 1007 power pack. This unit weighs less than 4 pounds and only takes up about 63 cubic inches of space. For operation between 9.0 and 10.0 volts, capacity between recharges would be about 80 hours. The initial cost of nearly forty dollars sounds pretty steep until you realize that the recharge cycle life of this unit is upwards of 500 cycles if care is taken to avoid over-discharge and if recharging is performed regularly. This represents over 40,000 hours of available power!

Recharging is easily accomplished with the aid of the Eveready 1807 charger, available for under ten dollars. Depending upon your store of ready cash and your eye for long-haul economy, this type of battery system could be the best selection.

lead-acid batteries

One other battery which should be mentioned is the familiar automotive-

type *lead-acid* battery. It has the advantage of being readily available and easily recharged for many, many cycles. Although it may be useful for some semi-portable or mobile applications, I would not seriously consider it for the example transceiver because of its objectionable bulk and messy liquid electrolyte.

making the choice

Looking at table 1, it is easy to see that the selection must be based on tradeoffs. The problem seems to reduce to choosing between six carbon-zinc D cells and the 10-volt nickel-cadmium power pack. In the case of the carbon-zinc D cells you get the lowest weight, least initial cost and a reasonable cost per hour. In the case of the nickel-cadmium power pack you get the smallest size and lowest extended cost per hour.

Which is best? That *depends on* your particular circumstances. For the transceiver let's choose the six carbon-zinc D cells, trading off increased bulk and higher long-term cost for the small initial cost. Table 1 also indicates that rechargeable alkaline D cells would be an attractive compromise selection.

The selection process is now almost complete. The friendly neighborhood drugstore has several different brands and prices from which to choose. If you pay a premium for the *Super Long-Lasting Heavy-Duty Power Cell* are you really getting a better cell than the cheaper *Brand X*?

To satisfy my own curiosity I bought two cells and conducted a simple test. Cell A was sold under the name of a popular discount department store at 34 cents for a pair. Cell B was a *Super Cell* manufactured by one of the well-known battery companies. It had 39 cents stamped on one end but was purchased in the same store for 48 cents per pair.

The test consisted of discharging the two cells into ten-ohm loads for about 3½ hours each day. Cell voltages were measured at the beginning and end of each discharge period and are displayed graphically in fig. 1. Note the recovery in cell potential after the cells are "rested." You can see from fig. 1 that cell B will last more than twice as long as cell A and is therefore the better buy in this case. In *general, it probably pays to buy the more* expensive batteries, even if only for the convenience of less frequent replacements.

conclusion

It should be obvious from this discussion that one of the requirements for intelligent battery selection is sufficient information on the available choices. Table 2 lists the names and addresses of several battery manufacturers from which further information may be obtained.

ham radio

table 2. Battery manufacturers. Most of these firms will provide information on the performance and recommended use of their batteries.

General Electric Co.
Battery Products Division
P.O. Box 114
Gainesville, Florida 32601

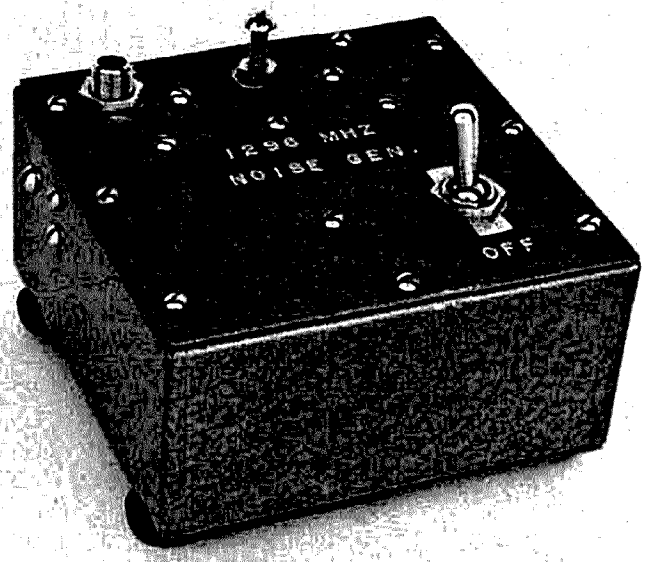
Gould, Inc.
Burgess Division
P.O. Box 3140
St. Paul, Minnesota 55165

Gulton Insustries, Inc.
Battery and Power Sources Division
212 Durham Avenue
Metuchen, New Jersey 08840

Mallory Battery Company
Broadway and Sunnyside Lane
Tarrytown, New York 10591

Sonotone Corp.
Battery Division
Saw Mill River Road
Elmsford, New York 10523

Union Carbide Corp.
Consumer Products Division 270 Park Avenue
New York, New York 10017



1296-MHz noise generator

Construction details for a simple diode noise generator for optimizing receiver performance on 1296 MHz

Cy Dirickson, W3BSV, 714 South Park Drive, Salisbury, Maryland

A noise generator is an extremely useful item to have around the shack for checking out your latest uhf preamplifiers or converters. This is particularly true on 1296 MHz where best low-noise performance must be obtained for practical, long-distance communications. The noise generator can also be used to check coaxial cable loss and transmit-receive relay loss by simply reading the values from your receiver's S-meter.

The output of the simple 1296-MHz noise generator described here is just

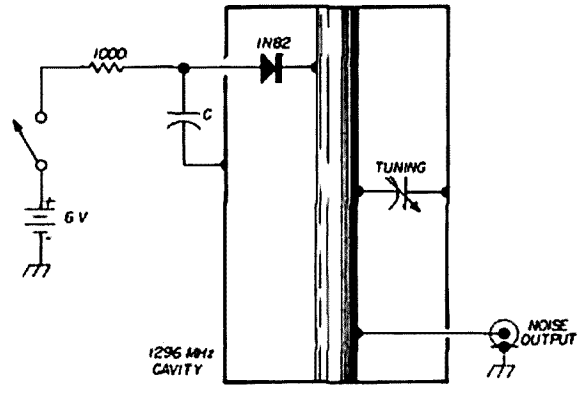


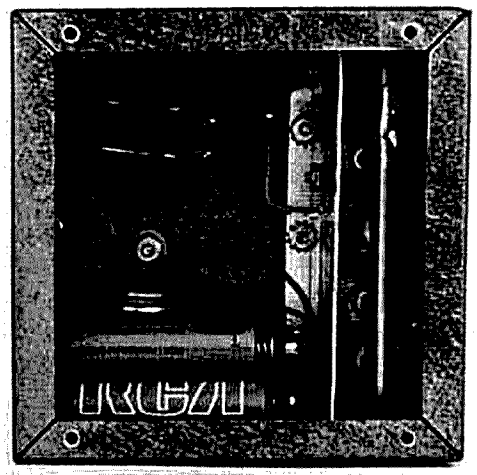
fig. 1. Simplified schematic of the 1296-MHz noise generator. Construction details are shown in fig. 2.

detectable on the average 1296-MHz mixer, and when fed through a K5200/K5500 rf amplifier, will show a gain of three S-units above internal noise. This represents approximately 18 dB.

construction

The 1296-MHz noise generator is built into a Bud 4x4x2-inch aluminum chassis with a bottom plate. All but one wall of the 1296-MHz trough line are provided by the chassis; the remaining partition is made from thin brass or copper sheet and attached to the chassis with 6-32 screws.

Both ends of the 1/4-inch brass rod used in the cavity are drilled and tapped for 6-32 screws. This rod is mounted in the center of the trough line with 6-32 screws, as shown in fig. 2. The 1N82 noise diode is tapped 1-inch down from the top of the line. The output BNC connector is connected to the bottom of the line with a 1/4-inch wide copper strap, 1-inch long. The copper strap runs parallel to the line as shown in the construction diagram.



Construction of the noise generator.

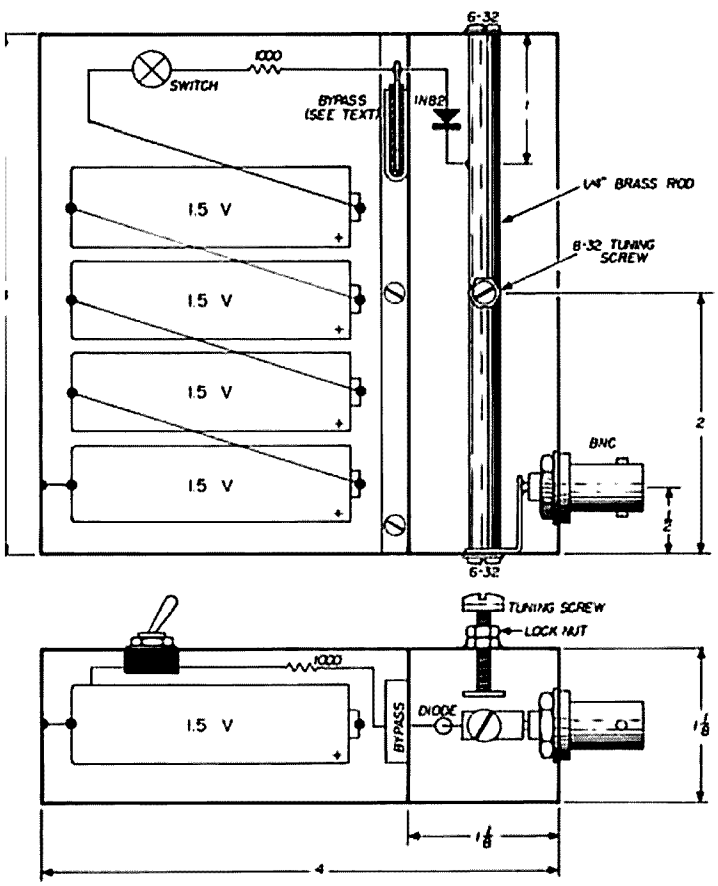


fig. 2. Construction of the 1296-MHz noise generator. Output BNC connector is actually located underneath the trough line, opposite the tuning screw, but is shown on the side, here, for clarity.

Although the BNC connector is shown on the side of the cavity for the purposes of this diagram, it is actually located underneath the trough line, directly opposite the 8-32 tuning screw.

The 8-32 tuning screw runs through a nut which is soldered to a thin, 2x3/4-inch brass plate which is attached to the aluminum chassis with screws. An 8-32 nut is used as the tuning disc. Another 8-32 nut is used for locking the tuning screw after the noise generator has been adjusted for maximum output at 1296 MHz.

The bypass capacitor consists of a 1-1/2x1-inch piece of thin brass, folded over into a U-shape and soldered to the side of the inside partition. Another small piece of brass sheet, about 1-inch square, is inserted into the U-shaped piece, as shown in fig. 2. Thin teflon sheet is used for the dielectric.

Power to the noise diode is provided by four 1-1/2-volt penlight cells, mounted in a compact battery holder. The spst toggle switch is mounted on the top of the enclosure.

ham radio

inexpensive all-channel frequency synthesizer for two-meter fm

How to obtain sixteen vhf fm channels, spaced 30 kHz, with only four inexpensive crystals

Jim Fahnestock, WØOA, Box 1355, Evergreen, Colorado 80439

One of the not-so-hidden expenses in two-meter fm operation is the cost of crystals. For 12-channels in some transceivers, the cost of crystals is nearly as much as the cost of the transceiver itself! And, on occasion, your desired operating frequency may *not* be among the 12 channels that cost so much to acquire.

The frequency synthesizer described here can produce 16 fm channels, spaced 30 kHz, with just *four* inexpensive crystals. With nine crystals all 36 channels from 145.98 to 147.03 can be synthesized. In addition, a five-crystal receiver local-oscillator synthesizer is suggested, covering all channels from 146.34 to 147.06 in 30-kHz steps.

The concept is quite simple, though different from (and much simpler than) conventional synthesizers. The idea has been *tried* on paper and appears to present no difficult problems. At the time of this writing a number of construction projects are in the works using ICs. One circuit uses four ICs, four crystals and two simple filters to produce 16 channels for repeater inputs. Another circuit uses eight crystals to provide 32 channels, permitting repeater *or* simplex operation on all commonly-used channels — and then some.

transmitter synthesizer

The basic block diagram of the transmitter synthesizer is shown in fig. 1. With

four crystals, all channels from 145.98 to 146.43 are covered, providing inputs for most repeaters with a transmitter that multiplies 24 times.

The upper half of table 1 shows the operating frequencies provided by the circuit of fig. 1. With five extra crystals, F_E through F_1, all channels from 145.98 to 147.03 are available in 30-kHz steps.

An alternate eight-crystal version is shown in fig. 2 and the available transmitting frequencies include all but 146.46 through 146.55 MHz. In fig. 2, the operating channels would be from 146.58 to 147.03 (16 channels) and the transmitter would be set to transmit *on channel* or 600 kHz below for repeater operation, simply by setting the *repeat simplex* switch in the appropriate position.

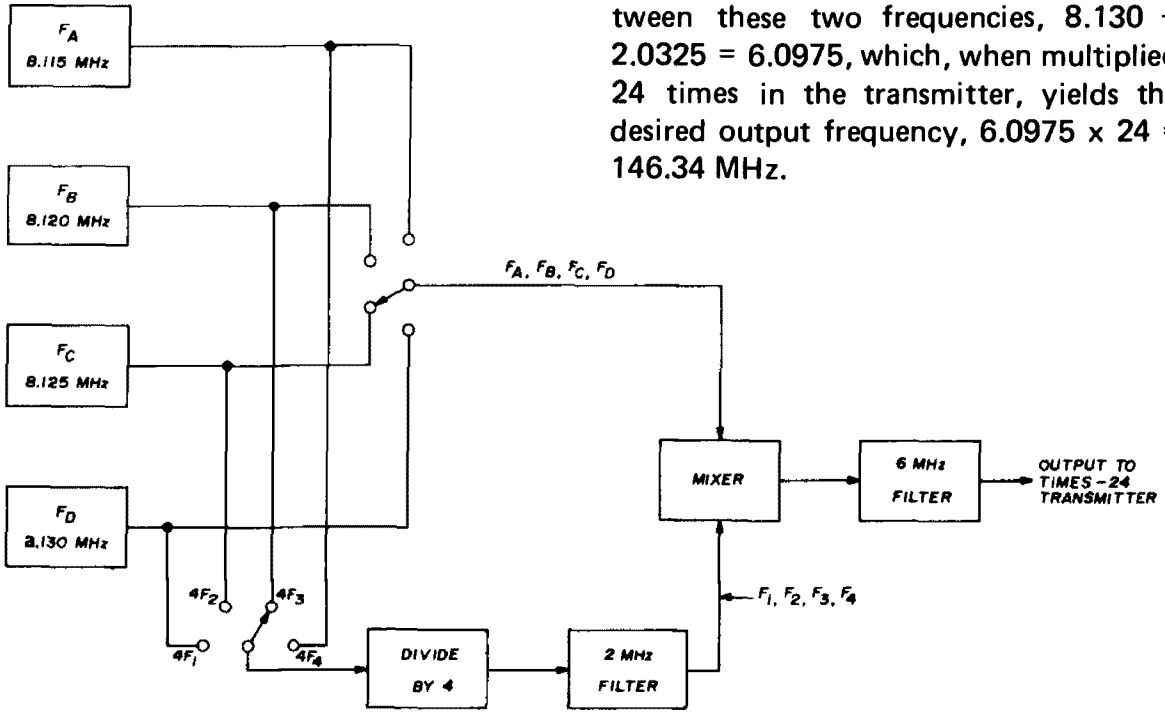


fig. 1. Suggested circuit for a 4-crystal, 16-frequency synthesizer to cover repeater input channels from 145.98 to 146.43 MHz. See table 1 for frequency combinations.

As shown in fig. 1, one crystal fundamental frequency is mixed with the fourth subharmonic (easily obtained with a single dual-JK flip-flop) of either the same or one of the other crystals to obtain the 24th subharmonic of the desired operating frequency. Multiplication is done in conventional fashion in the transmitter. The synthesizer simply serves as a substitute for the usual transmit crystals.

For example, suppose a frequency of 146.34 is desired. Referring to fig. 1 and table 1, this choice calls for mixing the fundamental of crystal D (F_D = 8.130 MHz) with the 4th subharmonic of crystal D (F_1 = 8.130/4 = 2.0325 MHz). The mixer extracts the *difference* between these two frequencies, 8.130 - 2.0325 = 6.0975, which, when multiplied 24 times in the transmitter, yields the desired output frequency, 6.0975 x 24 = 146.34 MHz.

The sixteen possible combinations of the four crystal frequencies and their 4th subharmonics provide the 16 repeater input channels as shown in table 1. Using 4th subharmonics of the same four crystals, but fundamentals of those shown in the simplex position, fig. 2, simplex operation is available.

Fortunately, 8-MHz crystals are inexpensive and easy to use. Garden variety crystals can be *rubbered* to the correct frequency in a variety of ways. If a high-frequency digital counter is available either the fundamental or the fourth

subharmonic of each crystal can be measured and adjusted. An accurate receiver or frequency meter capable of tuning 8 MHz or 2 MHz can also be used.

Notice that crystal D (F_D = 8130 kHz) can be set by the discriminator of a 146.34-MHz receiver since it is the only

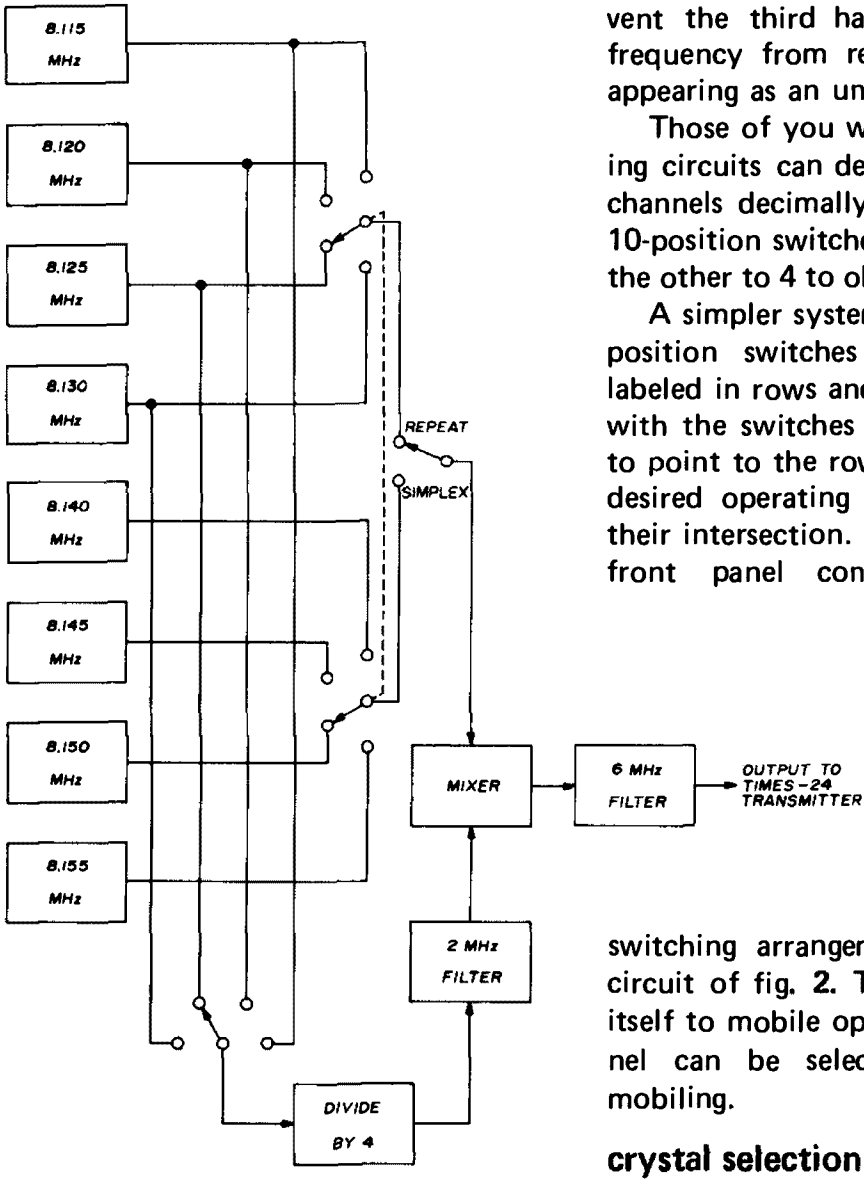


fig. 2. Circuit for repeater and simplex operation uses eight crystals for 32 transmit channels.

crystal used in generating that frequency. Likewise, crystals C, B and A can be adjusted with receivers on 146.25, 146.16 and 146.07 MHz. If the receivers are on frequency, all the other frequencies will be.

It should be noticed that all 16 frequencies are the result of either one or two crystals. This may require that two oscillators be operating simultaneously. Obviously, good isolation must be provided between oscillators to prevent the fundamental of the crystal whose 4th subharmonic is being used from reaching the mixer and producing a spurious signal.

The 2-MHz filter is necessary to prevent the third harmonic of the 2-MHz frequency from reaching the mixer and appearing as an unwanted 6-MHz output.

Those of you who are adept at switching circuits can design a system to select channels decimally, that is, to have two 10-position switches, setting one to 3 and the other to 4 to obtain 146.34 MHz.

A simpler system would use two four-position switches and a printed card labeled in rows and columns as in table 1 with the switches arranged mechanically to point to the row and column with the desired operating (receive) frequency at their intersection. Fig. 3 shows a possible front panel configuration using the switching arrangement suggested in the circuit of fig. 2. This arrangement lends itself to mobile operation since the channel can be selected by "feel" while mobiling.

crystal selection

At first glance, it might seem that the selection of crystal frequencies was somewhat arbitrary. However, there is just one set of frequencies that will give the desired operating frequencies and channel spacing for a given transmitter multiplication factor.

Several conditions must be met simultaneously. First, the *difference* between the selected fundamental frequencies and selected 4th subharmonic frequencies must be 1/24th the operating frequency

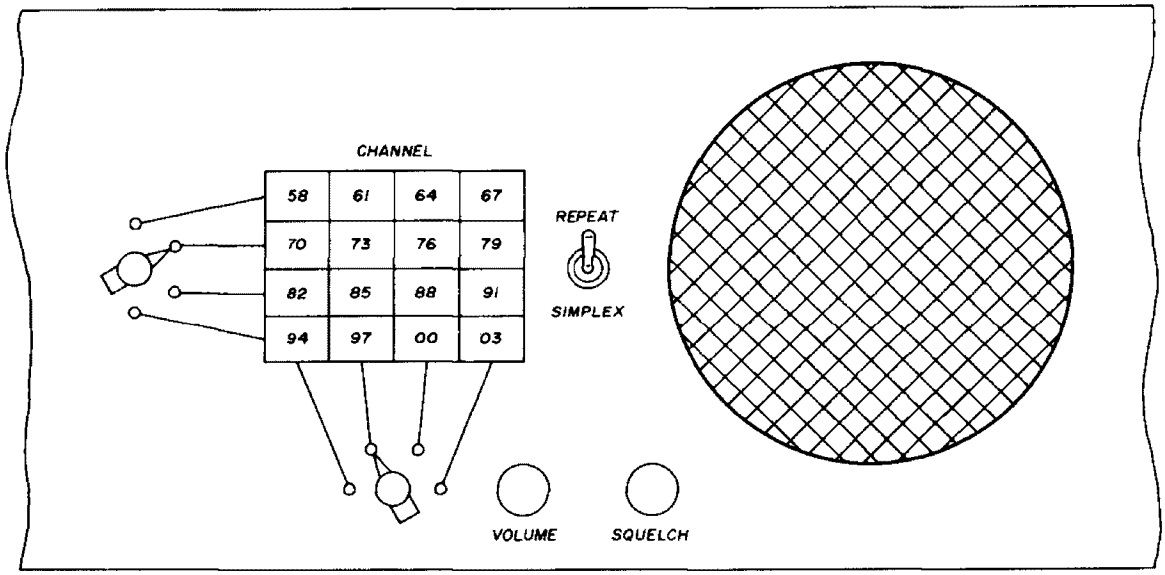


fig. 3. Suggested front panel layout for 32-channel transceiver shown in fig. 2.

for a 24 times transmitter. Second, the crystal subharmonic frequency *steps* must equal the desired channel spacing divided by the multiplication factor; since these frequencies are derived from the crystal fundamental frequencies, this determines the steps required of the crystal frequencies.

Since the goal is 30-kHz spacing in the 146-MHz region, the crystal subharmonic frequencies must be spaced 0.03/24 = 0.00125 MHz. The fundamentals must be spaced four times that, or 0.005 MHz.

The lowest operating frequency (in this case 145.98 MHz) will result from mixing the *lowest* crystal fundamental frequency with the *highest* subharmonic (the *difference* between the two determining the transmitting frequency). Crystal frequencies can be computed as follows:

$$F_A - \frac{F_D}{4} = \frac{\text{Lowest Frequency}}{\text{Multiplication Factor}} \quad (1)$$

Since the crystal fundamental frequencies must be spaced 0.005 MHz, and there are three *spaces* between F_A and F_D, then

$$F_D = F_A + 3 \times 0.005 = F_A + 0.015 \quad (2)$$

using this value for F_D in eq. 1 yields

$$F_A - \frac{(F_A + 0.015)}{4} = \frac{145.98}{24}$$

Solving for F_A, we obtain the 8.115

table 1. Transmitting frequencies for combinations of fundamental and 4th subharmonic frequencies. Four-crystal version covers channels above dashed lines. Eight-crystal version covers all channels except those between dashed lines. Nine-crystal version covers all channels from 145.98 to 147.04 MHz.

crystal frequencies	$F_1 = \frac{FD}{4} =$ 2.0325 MHz	$F_2 = \frac{FC}{4} =$ 2.03125 MHz	$F_3 = \frac{FB}{4} =$ 2.030 MHz	$F_4 = \frac{FA}{4} =$ 2.02875 MHz
F_A = 8.115 MHz	145.98	146.01	146.04	146.07
F_B = 8.120 MHz	146.10	146.13	146.16	146.19
F_C = 8.125 MHz	146.22	146.25	146.28	146.31
F_D = 8.130 MHz	146.34	146.37	146.40	146.43
		see text		
F_E = 8.135 MHz	146.46	146.49	146.52	146.55
F_F = 8.140 MHz	146.58	146.61	146.64	146.67
F_G = 8.145 MHz	146.70	146.73	146.76	146.79
F_H = 8.150 MHz	146.82	146.85	146.88	146.91
F_I = 8.155 MHz	146.94	146.97	147.00	147.03

MHz shown in table 1. Frequencies F_B, F_C and so on are obtained by successively adding 0.005 MHz for each fundamental frequency. To add the channels 146.46 through 146.55, a fundamental crystal of 8135 would be added with appropriate switching provisions.

It will be noticed that only subharmonics of F_A through F_D are used. You might be tempted to try a 5x5 matrix (or even a 6x6) to obtain additional channels with fewer crystals, but the frequencies become unwieldly and mixing problems arise. Getting 36 crystal-controlled channels with only 9 crystals isn't a bad compromise!

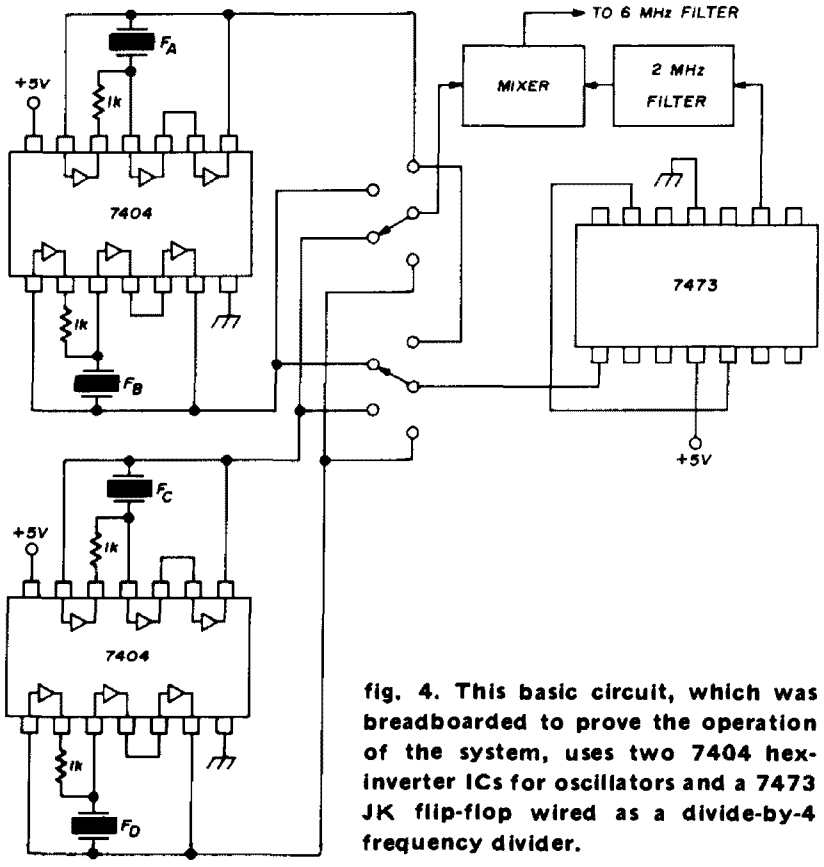


fig. 4. This basic circuit, which was breadboarded to prove the operation of the system, uses two 7404 hex-inverter ICs for oscillators and a 7473 JK flip-flop wired as a divide-by-4 frequency divider.

practical circuit

With no IC experience and even less equipment, I have managed to implement some of the functions illustrated in the foregoing block diagrams using readily-available and inexpensive ICs (see fig. 4). The 7404 hex inverter affords two oscillators in the circuit shown using series-resonant crystals of the values shown in table 1. No measurements have been made to determine the coupling between circuits. The 7473 dual-JK flip-flop makes an ideal divide-by-4 circuit and provides sufficient output to be heard in a Drake R4-A receiver equipped with a crystal for the 2.0- to 2.5-MHz band. Frequencies thus checked were right on.

The 2-MHz filter will probably consist of a parallel tuned circuit at 2 MHz and possibly a 6-MHz parallel circuit in series with the lead to the mixer to prevent the 3rd harmonic of the 2-MHz signal from reaching the mixer (the output of the divide-by-4 circuit is a square wave and will be rich in harmonics, suggesting careful attention to this potential source of spurious signals).

receiver frequency control

Naturally, the question arises, what about the receiver? There are several possible approaches. Probably the most simple is a tunable receiver (remember those?) with which the operator "finds" the repeater output or the 24th harmonic of a simplex frequency output, possibly with a *tune* position that activates only the synthesizer.

The circuit of fig. 5 illustrates another possibility. Here, the output of the synthesizer is multiplied 8 times (assuming a receiver with a local oscillator multiplication factor of three) and mixed with a crystal-controlled oscillator whose frequency is

$$F = \frac{F_{if} - F_{ro}}{3} \qquad (3)$$

Where F_{if} is the first intermediate frequency of the receiver and F_{ro} is the

table 2. Frequencies for receiver synthesizer using 10.7-MHz i-f and times-three receiver local-oscillator multiplication factor. Five crystals permit reception of 25 channels from 146.34 to 147.06 MHz.

	$F_1 = 5F_E$	$F_2 = 5F_D$	$F_3 = 5F_C$	$F_4 = 5F_B$	$F_5 = 5F_A$
	56.7667	56.7167	56.6667	56.6167	56.5667
$F_A = 11.3133$	147.06	146.91	146.76	146.61	146.46
$F_B = 11.3233$	147.03	146.88	146.73	146.58	146.43
$F_C = 11.3333$	147.00	146.85	146.70	146.55	146.40
$F_D = 11.3433$	146.97	146.82	146.67	146.52	146.37
$F_E = 11.3533$	146.94	146.79	146.64	146.49	146.34

repeater offset (usually 600 kHz). For the normal 600-kHz offset and a 10.7-MHz i-f, the frequency would be

$$F = \frac{10.7 - 0.6}{3} = 3.3667 \text{ MHz}$$

For simplex operation $F_{ro} = 0$ and $F = 10.7/3 = 3.5667$ MHz. Switching could be tied to the *repeat simplex* switch shown in figs. 2 and 3.

For the 34/94 situation, the synthesizer output of 6.0975 x 8 = 48.78 MHz, which, when mixed with 3.3667 MHz, yields a difference frequency of 45.4133 MHz. Multiplying this three times in the receiver results in the desired local oscillator frequency of 136.24 MHz which is 10.7 MHz below 146.94. The 45-MHz filter is necessary to eliminate the *sum* of the frequencies being mixed in the circuit.

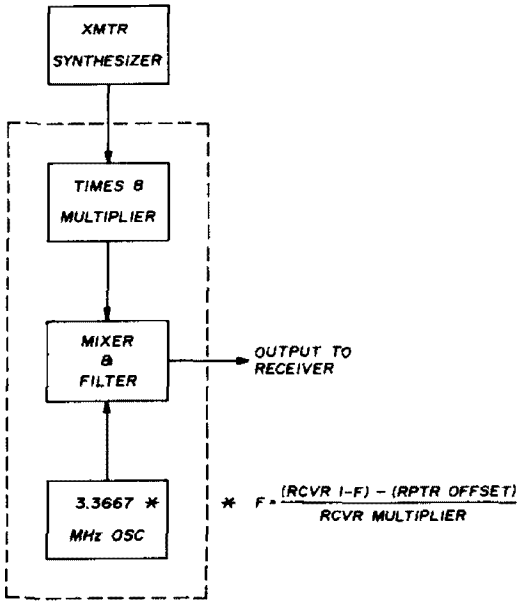


fig. 5. Tracking receiver local-oscillator frequencies can be obtained for repeater operation using this circuit. If reception is required on normal repeater input channels, 3.5667 MHz is substituted for the 3.3667 MHz frequency shown.

A slight modification of the matrix synthesizer described above can be used to generate local-oscillator frequencies for an all-channel receiver (see fig. 6). This technique lends itself to automatic scanning using a 5x5 matrix sequenced by two 5-bit shift registers.

Here, a 5x5 matrix is possible which generates local-oscillator frequencies for 25 receive channels from 146.34 to 147.06 MHz with only five crystals. Because the desired frequencies (to substitute for the receiver's first local-oscillator crystals) are higher in the receiver case, it becomes convenient to *multiply* the synthesizer crystal fundamental frequencies rather than to divide, as in the case of the transmitter synthesizer. The crystal formula is

$$F_1 - F_A = \frac{\text{highest frequency - receiver i-f}}{\text{receiver multiplier}} \quad (4)$$

but F_1 is the 5th harmonic of F_E, which must be 0.04 MHz higher than F_A for proper channel spacing (by the same reasoning used in the transmitter version) so $F_1 = 5F_E = 5(F_A + 0.04)$. Substituting in the above equation

$$F_A = 5(F_A + 0.04) - F_A$$

$$= \frac{147.06 - 10.7}{3} = 11.3135 \text{ MHz}$$

Other frequencies in the 5x5 receiver matrix are shown in table 2.

Here again, precautions must be taken

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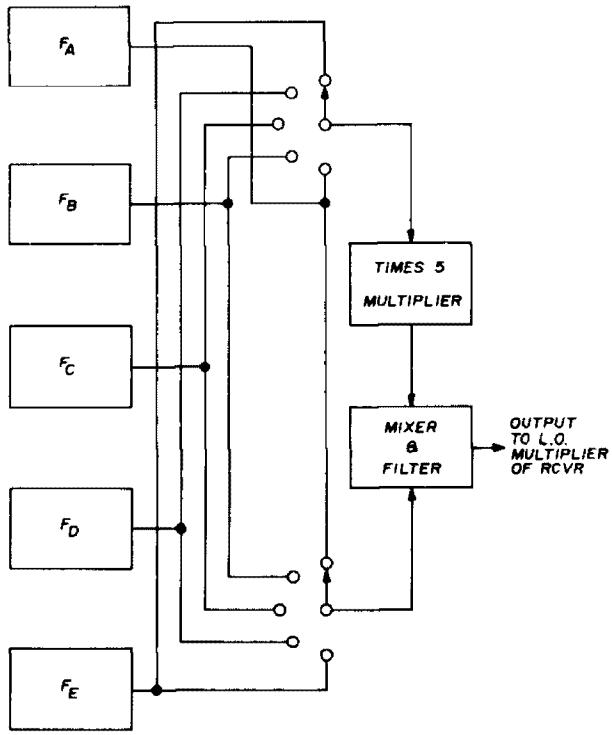


fig. 6. Suggested 5-crystal, 25-channel receiver local-oscillator synthesizer covering 146.70 through 146.34 MHz in 30-kHz steps. See table 2 for frequencies and crystal frequencies.

to minimize birdies and other spurious signals. For example, assume F_A and F_1 are being mixed to obtain the proper local-oscillator frequency to receive 147.06 MHz (56.7665 - 11.3113)/3 = 136.3896 MHz. It will be noticed that the 12th harmonic of 11.3133 MHz is 135.7596, which, if present in the receiver mixer, will admit a strong signal on 146.46. So, good filtering and isolation are important in designing a receiver around this matrix.

Although 10.7-MHz intermediate frequencies have been used as examples in this article, versions are possible for surplus equipment. Naturally, attention will have to be given to matching the outputs of these circuits to the requirements of the equipment to be used, but more difficult problems have been solved by ham ingenuity before and they should be solvable in this instance.

It is hoped that this idea article will inspire some interesting construction projects and promote more flexible and economical two-meter fm operation in general.

ham radio



comments

speech clipping

Dear HR:

I wish to make a few comments on G6XN's article on speech clipping. He states, "the clipping level should be sharply defined," but does not give a reason for this. Any nonlinearity in a system creates distortion and rounded corners on the envelope would create a low order product, increasing the possibility of the product falling within the passband. An audio or rf compressor should preceed the clipper to keep the clipping level constant as normal speech inflections (and emphasis on certain syllables) amount to 10 or 15 dB. How you can keep the clipping level constant without one is beyond me.

G6XN discusses filter shape characteristics rather thoroughly, but fails to mention that the second filter must have linear *phase* characteristics. Recall that a triangular wave (which has a lower peak to average than a sine wave) has the same harmonic content as the square wave we went to all the trouble to produce; you might even decide to use a maximally linear phase characteristic in the filter despite its lack of steep skirts.

I haven't tried it, but I think there would be more in-band distortion if anyone tried to use two-stage clipping with a third filter between the two filter stages. The reason for this is that the input to the second filter contains in-band distortion products created by the first clipper.

Mr. Moxon went through a lot of detail explaining why anything not essential should not be allowed to enter the clipper stage. This brings to mind an interesting possibility. Most speech information is contained in three ranges: 300-750 Hz, 1250-1750 Hz and 2250-2750 Hz (Technical Correspondance, *QST*, December, 1960). While the original article claims 10-dB improvement, I believe this to be for the receiving case as there is not that much speech power in the ranges 750-1250 Hz and 1750-2250 Hz.

With rf clipping it might be advantageous to build a three-band audio filter as the system is already rather complex and the filter system would not increase the complexity that much more. However, this must be tried as the results would be hard to analyze by analytical reasoning, as are most electronic systems involving the speech waveform. Anyone trying this should remember, however, that there should be provisions for switching the filters out if slow-scan tv operation is contemplated.

Larry Tessari, WA8ASD
Taylor, Michigan

In regard to the first paragraph, there seems to be some confusion between clipping level *which is fixed by, for example, the bias on a diode, and the* amount *of* clipping *which is the number of dB by which speech peaks exceed this bias level.*

It is important for the clipping level *to be clearly defined for the reason given on page 33 of my article. Any uncertainty about the peak level means that the drive to the final must be reduced by an amount equal to the uncertainty, since otherwise you cannot be sure of not overdriving the stage. On the other hand, my* fig. 1 *shows that the* amount of clipping *can vary between 15 and 25 dB for a little over 1-dB change in average signal level, so there is no need to keep this amount particularly constant.*

Nonlinearity is, of course, inseparable from clipping, but the experimental evidence is that any potentially harmful distortion is submerged under the wanted signal. Distortion is intermodulation not harmonic, and as those who have measured it say it is at least 10 dB down, it should amount to no more than a slight increase of background noise. The rounding of corners of the speech envelope is likely to be a function of the filter characteristic rather than the clipper. I have not found it necessary to use a compressor in front of the clipper but this would no doubt be an advantage to operators less used to speaking at a constant level.

It is an interesting question, however, as to what proportion of the statistical variations normally attributed to speech waveforms are inherent in the nature of speech and how much to inflections, emphasis, etc., occuring slowly enough to be taken care of by a compressor. This must obviously depend a lot on the individual.

Mr. Tessari raises an interesting point about the phase characteristics. I imagine the most serious consequence of a non-linear phase characteristic would be to delay high and low frequencies by different amounts, leading to amplitude variations at the output of the filter when both are present simultaneously. This could cause a system setup on i-f tones to go "over the top" occasionally on speech. I have not noticed this in practice but will have a more careful look for it.

It had not occurred to me to use a third filter between the two clipping circuits, and I would certainly agree that it would not be a good idea.

The 3-range idea is quite an old one, but my impression is that any restriction of bandwidth below that of an ordinary ssb filter produces a result which is not easy to listen to. This can be checked with a band-stop filter such as is sometimes used for removing CW interference. Rf clipping owes part of its effectiveness to boosting low-energy signals (if they occur in isolation) and this must presumably tend to fill in the 750-1250 Hz and 1750-2250 Hz gaps. I would agree, however, that it is not really possible to be analytical, and there are still plenty of unknowns in this field.

Leslie **A. Moxon, G6XN**

Dear **HR:**

I found the article by **G6XN** on the performance of rf speech clippers in the November, 1972, issue of *ham radio* to be quite informative and interesting. On page 33 the author refers to the question of acceptable passband ripple being halved by a two-filter system. I would like to point out that, based upon laboratory measurements, the passband ripple is actually *reduced* in relation to the amount of clipping, since it fills the valleys by nature of its basic process.

Typical current production filters, such as the Collins, etc., exhibit passband ripple in the area of a maximum 1.5 dB. More typically, in many that I have tested, they measure less than 1 dB.

Therefore, in rf clipping, with the conventional two-filter system, using well

matched filters for shape and impedance, no system degradation will be audible from the use of standard filters. This is confirmed by my own use of such a system for some time.

Marv Gonsior, W6VFR
Fullerton, California

rf speech clipping

Dear HR:

I read with great interest the paper on rf speech clipping by G6XN in the November, 1972, issue of *ham radio*. It is the best article I have read on the subject, and is a most informative and timely work.

I note that Mr. Moxon, on page 32, calls the theory behind pre-emphasis before speech clipping "obscure." On the contrary, I feel that it is quite straightforward as it applies to audio-frequency clipping.

One disadvantage of audio clipping, as compared to rf clipping, is that it generates in-band distortion products. These occur at odd-numbered multiples of the frequency of the clipped tone. For example, clipping a 100-Hz tone generates distortion products at 300, 500 and 700 Hz, etc.

In audio clipping, distortion products at 2500 Hz and higher are removed by a steep-skirted low-pass filter following the clipping stage. However, the distortion products of the lower frequencies are retained in the audio signal. These products are generated by the clipping of a tone below about 800 Hz. By minimizing the power in speech tones below 800 Hz, you can minimize the power of the in-band distortion products.

Unfortunately, however, much of the power density in normal speech is concentrated below 500 Hz, the most damaging region. In addition, these tones contribute little to communications intelligence. Therefore, they are best removed or attenuated before audio clipping.

Author OH2CD very effectively uses pre-emphasis in the clipper described in the February, 1972, issue of *ham radio*. In fact, he uses 12-dB per octave roll-off below 1400 Hz, and 18-dB per octave roll-off below 300 Hz.

The argument concerning distortion products does not apply to rf clipping, but pre-emphasis may nonetheless be useful since energy in speech tones below 500 Hz contributes little to communications intelligence while consuming considerable amounts of transmitter power.

Tom Ashley, WB4SIJ
Lexington, Kentucky

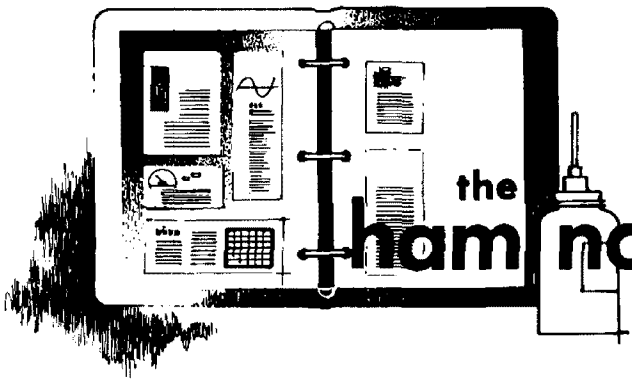
technical education

Dear HR:

I read with interest "A Second Look" in the January, 1973, issue of *ham radio*. Your observations on the trend in technical education are, in my opinion, accurate and timely. The collegiate level engineering studies have indeed reached a degree of sophistication that renders the graduate incapable of performing at the job site. These institutions, in striving for academic excellence, have failed to establish priorities relative to the actual industrial needs.

The real need in our field today is, as you stated, a "gut" understanding of electronics rather than mathematical sophistication. An alternative that you did not mention (and one that could perhaps contribute to a future article) is the practical and applied curricula offered by the many career training institutes around the country. These schools stick to the "gut" material and supply a large portion of the qualified technical personnel now being used. An excellent source of information is the National Association of Trade and Technical Schools, 2021 L Street, NW, Washington, D.C.; William A. Goddard, Executive Director.

E. R. Massengill, President
Tennessee Institute of Electronics
Knoxville, Tennessee



the ham notebook

sideband location

Frequently, you can find 30 to 50 upper-sideband stations in one day, operating within one kHz of 14349 or 21449 kHz, thus with the sideband extending over the band edge. This is contrary to FCC regulations, Section 97.63. In addition, Section 97.75 requires independent means for measuring the carrier frequencies, accurate enough to ensure keeping within the band. A bit of listening discloses the *fact that many amateurs are* uninformed, or are unwilling to comply.

Fig. 1 is an approximation of the situation in the Collins 32S-3 exciter. The carrier is off to one side so that it is suppressed 30 dB because of that, and *another 20 dB or so in the* balanced modulator. Using only the instruction manual information, not supported by measurements, you will note that the 2100-Hz filter width, 6 dB down, actually extends more than 2100 Hz above the carrier frequency because of the position of the carrier down on the side of the filter curve. At 30 dB down, the voice extends 2700 Hz above the carrier frequency.

In the Swan 500, fig. 2, even at 6 dB down, the modulation extends more than 2700 Hz higher in frequency. At 20 dB down, the voice extends to 3500 Hz above the carrier. At 60 dB down, the filter width is given as 4400 Hz, but the published curve is about 5 kHz wide there, of which about 1 kHz is below the carrier and some 4 kHz is above.

Filters may not all be as symmetrical as these. Some difference in the shape on one or the other side of the curve may be found.

Not many amateurs appreciate that the military and commercial frequency assignments are for the *midpoint* of the emission, *not* the suppressed carrier frequency which is widely used in amateur

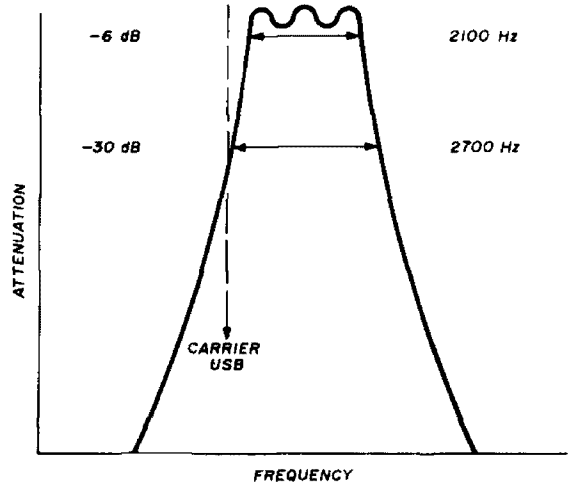


fig. 1. Bandwidths for Collins 32S-3 exciter in the USB mode.

radio. As a result, a MARS transmission in USB on a dial setting of 20998 kHz normally will extend appreciably into the bottom of the 15-meter band.

In order to keep the MARS stations on the right frequency, Navy-Marine Corps MARS *has* published **fig.** 3 in their instructions, DNC 8 (A). This is somewhat simplified, in that it does not stress the separation of the suppressed carrier from the modulation, nor any level of attenuation at the edges of the modulation. The illustration shows a presumed 3-kHz voice channel extending up from

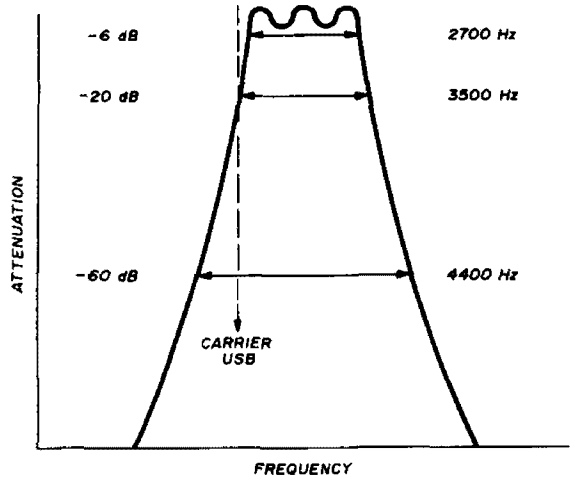


fig. 2. Bandwidths for Swan 500 transceiver in the USB mode.

the suppressed carrier. Therefore, you must set the exciter's suppressed carrier 1.5 kHz *below* the assigned frequency for USB transmission. As will be seen from fig. 2, some equipment would not keep within a 3 kHz channel unless you could accept perhaps only 12 dB attenuation outside of the channel; and the optimum detuning of the suppressed carrier would be slightly different.

From the illustrations it can be seen that fairly reasonable attenuation results when the Collins 32S-3 is operated in USB mode 3 kHz inside of the upper band edge, whereas the Swan 500 should stay inside by at least 4 kHz, and then only with suitable calibration equipment to ensure accuracy of the frequency measurement. A linear amplifier will raise the level of the emission and, therefore, of the filter curve, thus inviting greater interference to adjacent channels.

Minimum power is required by the Communication Act, Section 324, now written into the Amateur rules, Section 97.67 (b), as quoted in the October, 1972, issue of *QST*, page 108. Unnecessary power may create considerable unnecessary splatter due to the filter skirts, equipment problems or nonlinearity in the several stages following the ssb filter and on through the linear amplifier. In addition, there may be off-frequency emissions greater than those indicated by the filter skirts demonstrated by the above tests. Furthermore, the transmitter can have a somewhat different filter performance than the receiver.

Bill Conklin, K6KA

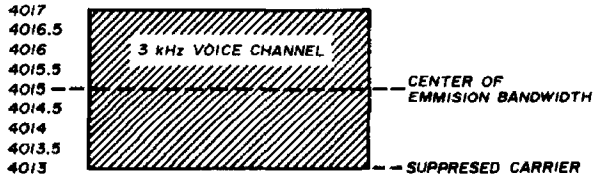


fig. 3. Example of emission published in Navy-Marine Corps MARS Communication Instructions, DNC 8 (A).

swr bridge readings

When you can't obtain full-scale deflection on your swr bridge because of low power output or low antenna impedance, you may still determine the swr with fair accuracy by using the half-scale values listed in table 1.

table 1. Typical swr values when swr bridge is adjusted for half-scale deflection on forward power instead of full-scale deflection. This may be the case if power output is low or the output frequency is high.

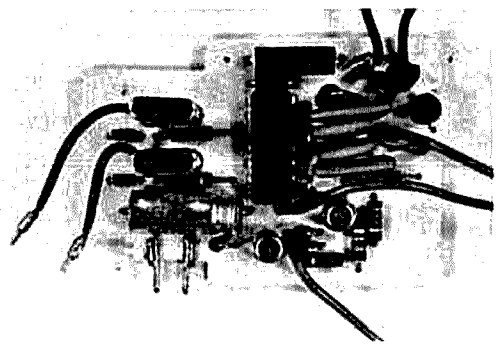
half-scale swr reading	actual swr
1.1	1.2
1.2	1.3
1.3	1.5
1.4	1.7
1.5	2.0
1.75	3.0
2.0	3.5
2.5	4.0
3.0	infinite

The swr bridge is calibrated in terms of half-scale forward meter deflection instead of full scale deflection. First, adjust the bridge for full-scale deflection in the forward direction. Produce a high swr reading (3:1 or 4:1) by detuning the antenna coupler or by any other convenient means. Now, reduce the forward reading to half scale *and note the swr reading*. Repeat the above steps until you have tabulated a complete set of swr readings as shown in table 1.

F.J. Bauer, W6FPO

new products

vhf fm scanner



Topeka Engineering has just announced a two-channel frequency scanner for the Regency HR2 and HR2A transceivers, the *Scan 2*. The unit plugs into the HR2 with only three solder connections and one component change. The scanner unit mounts internally above the A, B and C crystal sockets. Scanning is accomplished by inserting the desired crystals into positions A and B. A and B channels will then be scanned when the switch is in position C so no change is required to the radio.

A unique search-back feature of the *Scan 2* allows the unit to scan a channel not in use every 5 seconds; this may be disabled to switched in or out. The scan rate is 20 times per second when no channel is in use. The *Scan 2* is available at $19.95 including shipping from Topeka FM Communications and Electronics, 1313 East 18th Terrace, Topeka, Kansas 66607. For more information, use *check-off* on page 94.

midland enters amateur radio market



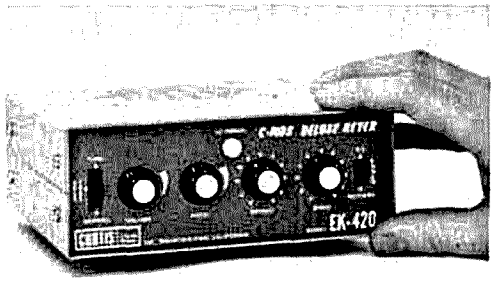
Midland Electronics Company, a leader in communications equipment for over a decade, has entered the amateur radio field. Leading off Midland's amateur radio offerings is a 15-watt, 12-channel, 2-meter fm mobile transceiver. This is a compact unit, 2-1/4" high by 6-3/8" wide by 8-7/8" deep, weighing only 4-1/2 pounds. It provides 15 watts rf output (1 watt in the low-power position) and contains a multiple fet front end receiver with high-Q helical resonator filters and ceramic filters. The model 13-500 operates on 12 volts dc and is supplied with push-to-talk microphone, mobile-mounting bracket and crystals for 16/76, 34/94 and 94/94. Suggested retail price is $249.95. Additional crystals covering the full 2-meter range are available from Midland.

Midland's 10-watt, 12-channel 220 fm unit brings a much-wanted, low-cost mobile transceiver to this growing amateur band. The model 13-509 transmits at 10 watts or 1 watt rf output, and has the same compact size and light weight of the two-meter mobile equipment. Features and equipment are similar throughout and 223.00 simplex crystals are supplied. Suggested retail price is $219.95.

All of these new Midland amateur radio transceivers are described fully in a colorful new catalog. For a free copy of this catalog or information on Midland's amateur radio dealer franchise program, contact the Midland Sales Manager, Midland Electronics Company, Box 19032, Kansas City, Missouri 64141, telephone (816) 474-5080, or use *check-off* on page 94.

More Details? CHECK–OFF Page 94

electronic keyer



A super low-power electronic keyer using complementary MOS (C-MOS) integrated circuits has been announced by Curtis Electro Devices. For QRP in the field or the full gallon at home, the EK-420 C-MOS Deluxe Keyer offers effortless, self-completing dots, dashes and spaces, dot memory, iambic operation, built-in side-tone and 4-inch speaker, built-in power supply and a reed relay for grid-block keying or solid-state rigs. It operates on 115 Vac or +4.5 to +14 Vdc.

Because of C-MOS circuitry, the EK-420 consumes only ½ mW during standby. When keying, power consumption rises to 180 mW. This allows extended operation on an inexpensive snap-in 9-volt transistor battery. A rear-panel accessory socket allows instant connection of a soon-to-be-announced four program, 2048-bit (200-character) memory unit.

Front-panel controls include sidetone volume and pitch, weight and speed, tune, self-test, and ac-battery selector switch. All eight ICs, three transistors and the reed relay plug in to allow easy field service.

Housed in a heavy gauge case, the EK-420 is priced at $139.95 complete. For more information contact Curtis Electro Devices, Inc., Box 4090, Mountain View, California 94040, or use *check-off* on page 94.

mobile antenna mount

The new Larsen Electronics mobile antenna mount, described in the June, 1973, issue of *ham radio*, is designed for the usual 3/4-inch hole, *not* a 3-inch hole. The mount is available from Larsen Electronics, Inc., Box 1686, Vancouver, Washington 98663.

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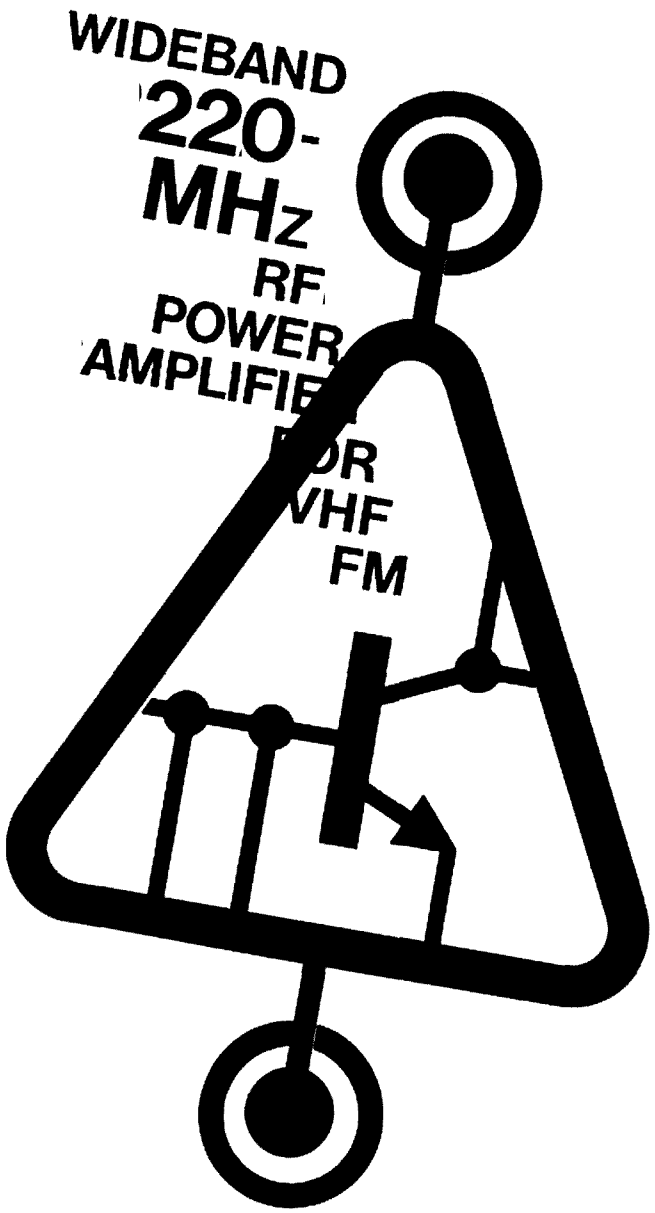
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ham radio magazine

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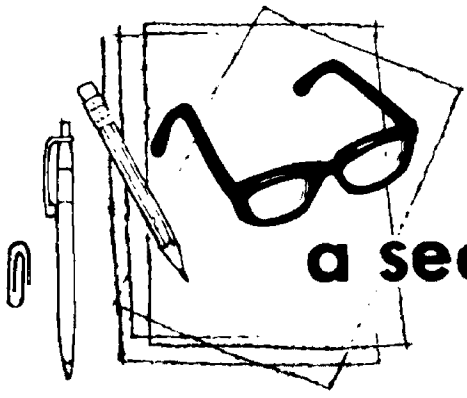
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a second look

by jim fisk

circuit design contest

Here is an opportunity for all you experimenters and amateur circuit designers to put your talents to work, and perhaps, to win some fabulous prizes. This contest which has the theme, "Quadzilla is coming", is sponsored by National Semiconductor and *EDN* magazine, and is for the best circuit design using all four devices in *any* linear quad IC. National Semiconductor will pick the 10 winners, while *EDN* will determine the placement of those winners.

Although anyone can enter the contest, the entry must be on a National Semiconductor entry blank or a facsimile thereof. National is offering a 32-page handbook on the contest which is half data sheets on National Semiconductor Quad ICs, and half application notes, and includes an official entry blank. Copies of this useful handbook are available, upon request, from *ham radio* magazine.

The grand prize winner will receive a Panasonic 4-Channel Entertainment system which includes a tape and record player, receiver and speakers. The next four winners will receive a Zenith portable color TV set. The last five winners will receive Garrett Incredible Time Machines — electronic digital calculators with a built-in digital alarm clock. Entries must be submitted by October 31st, so now is a good time to dust off your slide rule and put your ingenuity to work.

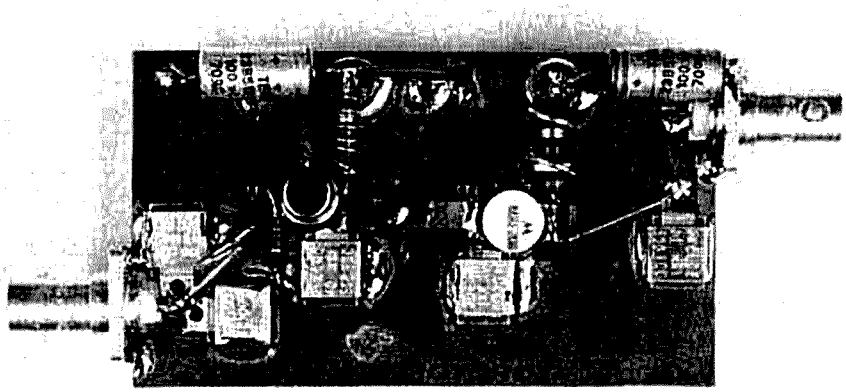
220-MHz repeater

In an effort to bring vigorous 220-MHz activity to amateurs from coast to coast, Clegg has just announced a repeater-lease program, where a new Clegg 220-MHz repeater, valued at approximately $1200, will be leased to amateurs at special club rates of only $25 per month. Although FCC Docket 19759 proposes to convert the top 1-MHz of the 220-MHz amateur band into a new class-E Citizens Band, it may still be possible to prevent adoption of this proposal if it can be demonstrated that large numbers of amateurs are using the 220-MHz band. The new Clegg 220 repeater-lease system appears to provide just such an opportunity.

The 220 repeater is leased complete (except antenna and transmission line), with features that include automatic identification, all solid-state construction, and built-in timers. It operates at 10- to 15-watts output, uses a Phelps-Dodge duplexer, and has approximately 0.4-μV sensitivity. The repeater includes an ac power supply, local microphone input and metered signal strength.

All amateur radio clubs are invited to contact the Clegg Division if they are interested in getting their club into this repeater program. The low monthly rental fee can be further reduced with club member purchases of the Clegg FM-21 transceiver, a 220-MHz fm unit. Interested clubs should write to Phil Theis, K3TUF, Clegg Division, International Signal and Control Corporation, 3050 Hempland Road, Lancaster, Pennsylvania 17601, or telephone him at (717) 299-3671 for more information.

Jim Fisk, W1DTY
editor



220-MHz rf power amplifier for vhf fm

Design and construction of a solid-state, 10-watt, class C rf power amplifier with an operating bandwidth of 40 MHz

Jerry DuBois, K7JUE, Motorola Semiconductor Products, Inc., Phoenix, Arizona

Designing high-gain, wide bandwidth, vhf power amplifiers has been a sticky problem in the past. However, by using a new approach, and standard rf semiconductors, you can now get improved performance at a lower cost. The particular rf amplifier described in this article was designed to operate at 220 MHz with an output power of 10 watts and a bandwidth of about 40 MHz.

Two members of a new high-gain family of Motorola vhf devices are used here to demonstrate their effectiveness as broadband amplifiers. The MRF207 and MRF208 rf power transistors, when used together, net greater than 100 times power gain at 220 MHz.*

The design goals included successful broadband vhf design — something which has often eluded rf designers. At the same time, a high-gain amplifier was needed which would supply 11-watts output with a reliable 20-dB of stable gain. With a low-pass filter at the output this would provide a solid, 10-watt radio. A block diagram of a transmitter using this amplifier is shown in fig. 1.

*The Motorola MRF207, MRF208 and MRF209 are available from franchised Motorola distributors. In small quantities the MRF207 is priced at $2.25, the MRF208 at $11.70, and the MRF209 at $14.25.

Recent developments in low-Q transistors have stimulated new approaches to wideband design. With the publication of complete data sheets, trial and error has been virtually eliminated from matching network design. In the following design, real life values are within a few percent of their calculated counterparts.

sider the transistor chip capacitance as the first element.

Referring to the Smith chart analysis shown in fig. 2, the input impedance of the MRF207 (10 - j11.5 ohms) falls in the upper lefthand corner of the chart. (This chart is normalized to 1, so all impedance values on the chart must be multiplied by

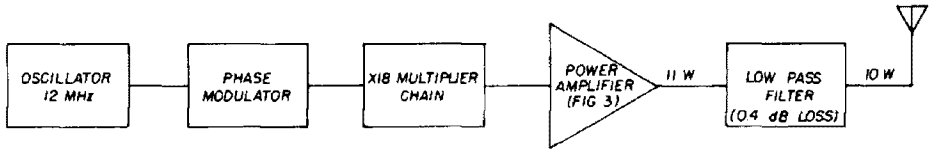


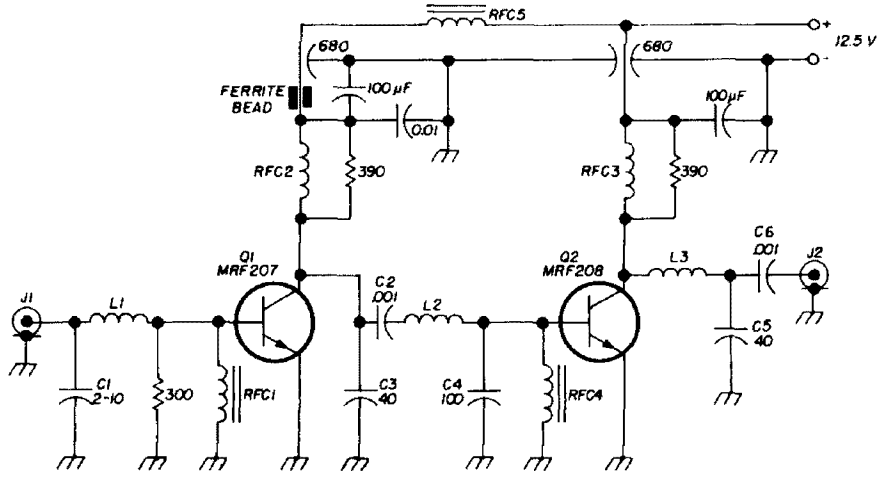
fig. 1. Typical 10-watt, 220-MHz fm transmitter using the rf power amplifier shown in fig. 3.

design

Achieving gain flatness, along with good efficiency, involves some trade-offs. The trick is to know when to stop. The amplifier described here displays good efficiency characteristics over the entire design bandwidth with a gain variation of about -0.8 dB. In this particular circuit, single step networks, which are broadband by nature, are used. Actually, they represent a pi configuration if you con-

50, all admittance values must be multiplied by 20.) First, a *series* element is plotted which transforms 10 - j11.5 ohms to 10 + j20 ohms. The length of this series rotation is 0.61 ohms on the chart; multiplying by 50 yields 31.5 ohms. A reactance of +31.5 ohms at 220 MHz is provided by a 22-nH inductor.

To transform the 10 + j20 point to 50 + j0 requires a *parallel* element. Since parallel circuit elements are most easily



C1	20-pF metal-clad mica capacitor (EI Menco MCM 01/002/-CA200DO)
C2,C6	0.001-µF metal-clad mica capacitor (EI Menco MCM 01/002/-CA103DO)
C3,C5	40-pF metal-clad mica capacitor (EI Menco MCM 01/002/-CA400DO)
C6	100-pF metal-clad mica capacitor (EI Menco MCM 01/002/-CA102DO)
L1	1 turn number 24 wire, 1/4" ID
L2	copper strap, 0.032" thick, 0.25" wide x 0.75" long
L3	0.8" lead of capacitor C6 (0.001-µF disc)
RFC1,RFC4 RFC5	low-Q rf choke (Ferroxcube VK200-20/4B)
RFC2	2 turns no. 24 wound around 390-ohm, 1/4-watt resistor
RFC3	2 turns no. 20 wound around 390-ohm, 1-watt resistor

fig. 3. Schematic for the 10-watt, 220-MHz rf amplifier. Capacitors C1, C2, C3, C4, C5 and C6 must be metal-clad mica types such as those manufactured by EI Menco. Other types of capacitors will not result in proper circuit operation.

plotted when described in terms of admittance, first convert 10 + j20 ohms (0.2 + j0.4 on the chart) to 20 - j40 millimhos (1.0 - j2.0 on the chart). The length of the parallel rotation to the center 50-ohm point is 40 millimhos or 25 ohms, represented at 220 MHz by a 20-pF capacitor.

If you are able to use an immittance chart you can eliminate the impedance to admittance conversion step described above, since it is accomplished automatically by the immittance chart. However, the Smith chart is somewhat more familiar to amateurs, hence its use here.

This simple procedure completes the design of the input network. The other two networks are designed in the same fashion (see fig. 3). The interstage network is really a two-step network with the device input inductive reactance acting as one element. Series equivalent impedance data for both transistors is given in table 1 so the designer can see how all network element values were arrived at.

It is obvious that this is a single-frequency design. Since the device characteristics vary gradually with frequency, these transistors will remain matched over a relatively large frequency range. Multi-frequency data could be treated in the same manner, and a network which has flatter frequency response might be calculated. However, in this case, the 40-MHz

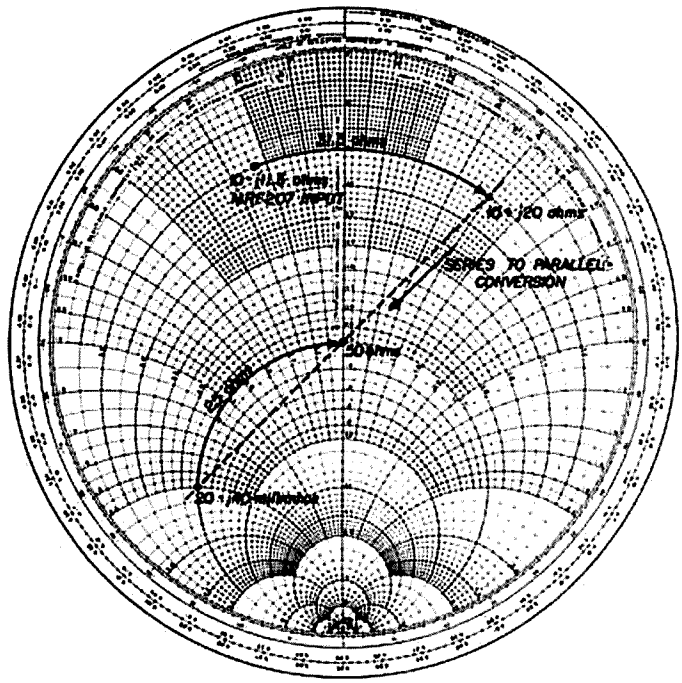


fig. 2. Smith chart analysis of the input network used to match 50 ohms to the series equivalent input impedance of the MRF207 transistor, 10 - j11.5 ohms.

bandwidth was considered to be more than adequate when the entire 220-MHz fm band is only 2-MHz wide. The additional work and increase in component count was deemed unnecessary.

construction

Construction of the 220-MHz power amplifier is generally straightforward, but some areas require care. The decoupling networks are important to stability.

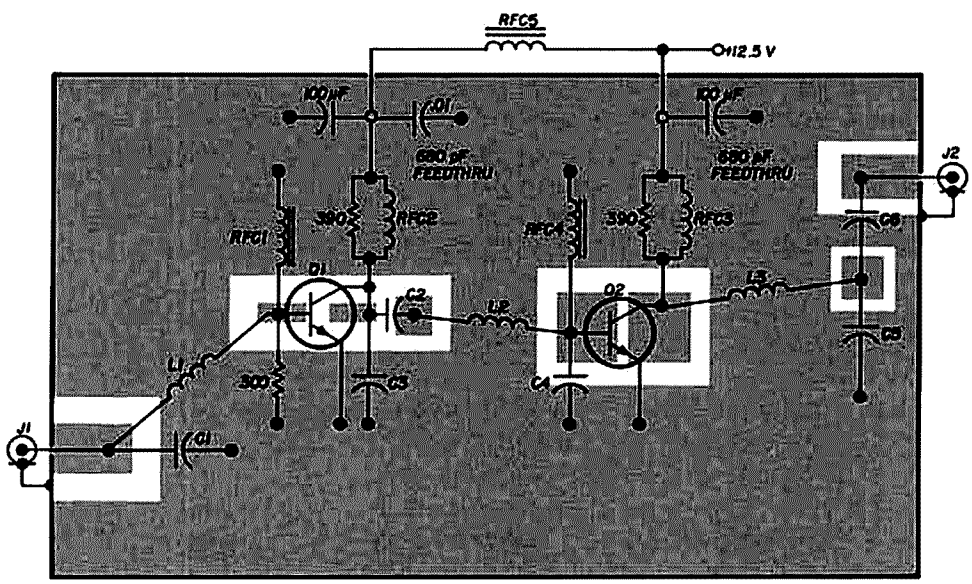


fig. 4. Full-sized printed-circuit layout for the 10-watt 220-MHz power amplifier.

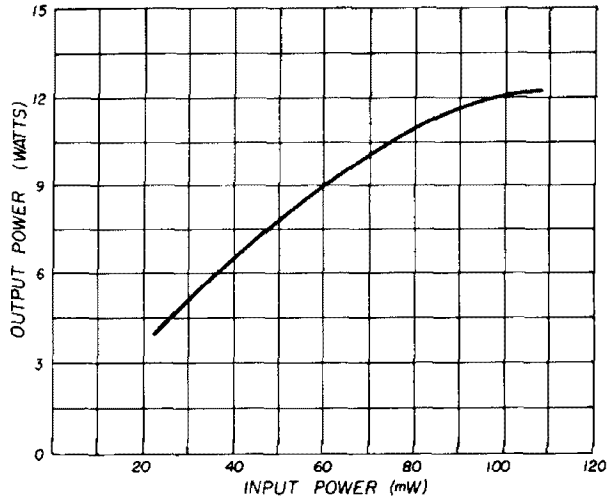


fig. 5. Power output vs drive for the 220-MHz power amplifier operating with a 12.5-volt power supply.

Feed-through type construction was found to be necessary. One matching element, L2, may require adjustment. Tweaking can be accomplished by varying the point on L2 where the 0.001-μF blocking capacitor is connected. This capacitor was purposely used as part of the network to allow adjustment. A good place to start is about 2/3 of the way out from the base of the MRF208 (see photograph). Inductor L2 must be soldered very close to the case of the MRF208. The 100-pF Unelco capacitor, C4, must also be mounted very close to the transistor as shown in the photo.

Inductor L3 can be increased in value slightly to give better efficiency at some sacrifice in bandwidth. As with any high-gain vhf circuit, minimum emitter lead length is required for maximum gain and bandwidth. Inductors L2 and L3 may be bent closer to or away from the printed-circuit board to tune the amplifier.

A clip-on heatsink should be used on the MRF207 driver transistor, and a heatsink approximately 2x2x0.5 inches should be used with the MRF208 output transistor. A full-size printed-circuit lay-out for the amplifier is shown in fig. 4.

A gain curve is included in fig. 5 which shows that the amplifier has some considerable reserve at a specified power input of 100 mW and 10-watts output. The input vswr varies from 2:1 at the band edges to 1:1 at 220 MHz. Plots of worst-case amplifier performance show a total output variation of 0.8-dB over the frequency range, 200-240 MHz (see fig. 6). Efficiency runs approximately 55% and spurious outputs are down more than 35 dB.

table 1. Series equivalent impedance data for the Motorola MRF207, MRF208 and MRF209 rf power transistors at 220 MHz.

device	power rating (watts)	input impedance (ohms)	output impedance (ohms)	minimum gain at 220 MHz (dB)
MRF207	1.0	10 - j11.5	32 - j41	8.2
MRF208	10	1.4 + j1.4	5.7 - j1.25	10.0
MRF209	25	1.4 + j1.8	3.9 - j0.2	4.4

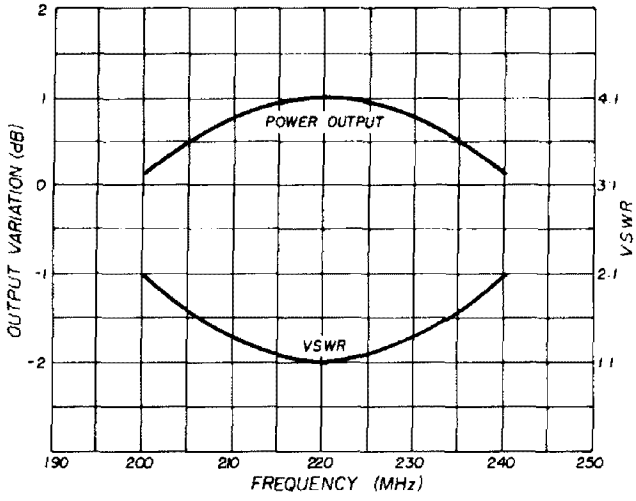
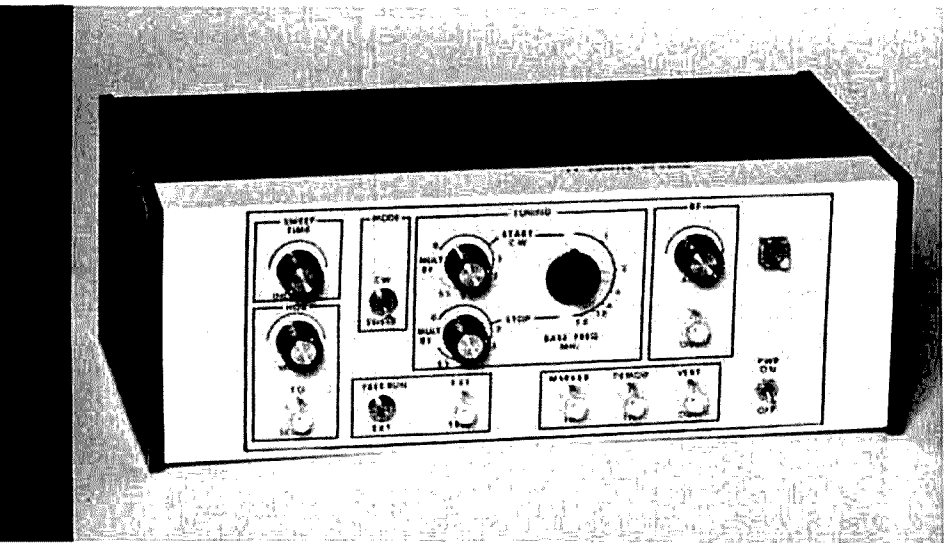


fig. 6. Power output and vswr characteristics (50-ohm load) for the wideband power amplifier. Gain varies less than 0.8 dB from 200 to 240 MHz.

This 220-MHz power amplifier shows that a wideband amplifier in the vhf region, built with currently available, non-internally matched transistors, is feasible. In this amplifier we have replaced an old design which had six tuning adjustments with a fixed-tuned power amplifier that has 10 times the bandwidth. At the same time, a reduction in both cost and component count has been accomplished.

ham radio



solid-state i-f sweep generator

Complete construction details for an i-f sweep generator that covers the frequency range from 100 kHz to 15 MHz

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It has been said that the things we see are more believable than effects perceived by any of the other senses. Also, a picture can convey more information in a moment than other forms of measurement can over a much longer period of time. Undoubtedly, this reasoning is responsible for the wide range of sweeping-type electronic instruments in common use today.

The sweep generator described here is similar to one I discussed in a previous article.[1] While more simple and less costly to build, it boasts some worthwhile improvements over its predecessor and should not be relegated to the gadget category. Keep in mind that this design is quite flexible and lends itself to innovation. Some of the possible variations will become evident later in the article.

the vco

The heart of any sweeping instrument is generally a vco. In this case a Motorola MC4024P IC multivibrator (vcm) is used

as the signal source. Square waves may not be ideal for rf but they do work, and this economical IC provides good sweep capability too. It also eliminates the need for alc circuits since the output amplitude is constant.

The MC4024P contains two identical oscillators and level translators but only one of them is needed in this application. The only external component required is a capacitor, which, in combination with a dc control voltage, determines the oscillator frequency. Thus, two methods for determining the frequency range of the oscillator suggest themselves. Use several fixed capacitors in conjunction with a range switch or a single high capacitance variable to cover the desired range, as I did here.

In either case, the selected capacitance determines what shall be referred to as the "base frequency." The dc control voltage will then establish the extent of the sweep excursion.

Voltage control of the oscillator frequency is accomplished by applying a voltage ramp to the proper pin on the IC. The dc levels at both ends of the ramp are set by front-panel controls, thus allowing independent selection of both the starting frequency and the stopping frequency.

Although the data sheet for the MC4024P IC claims that 3.5 to 1 variation in frequency is possible with control voltages ranging from 1.0 to 5.0 volts, all the devices I tried provided at least 5 to 1 change and as high as 6 to 1 if the control voltage was increased to about 6.0 volts.

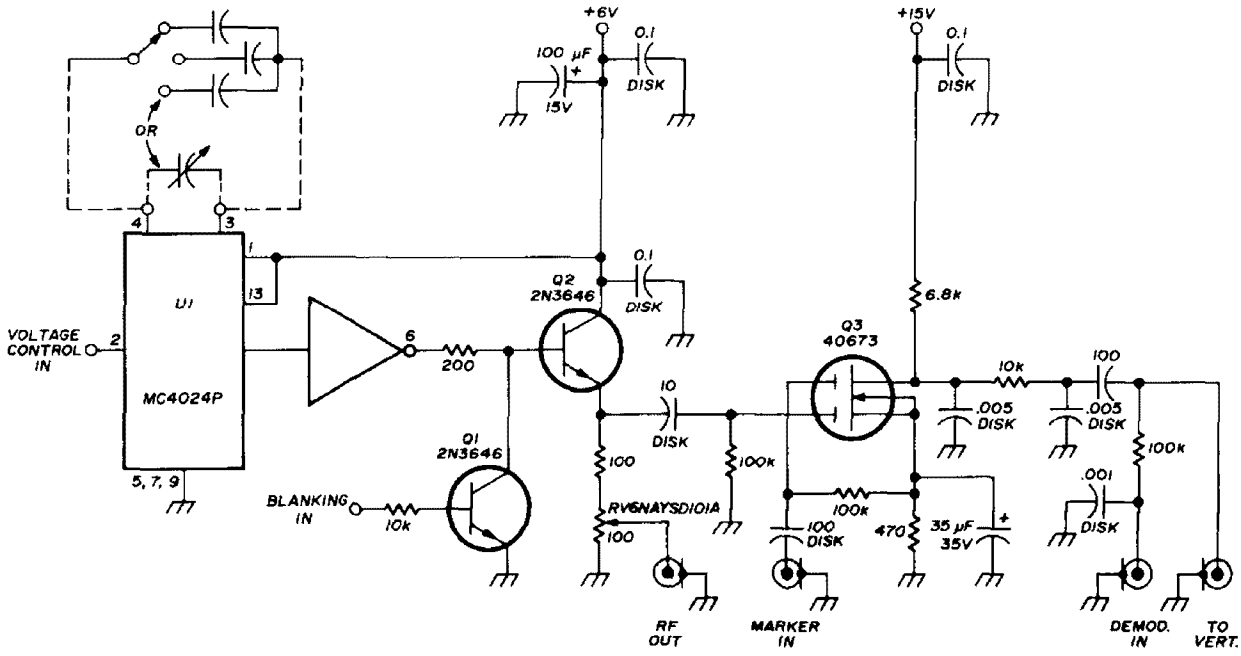


fig. 1. Rf oscillator and birdie marker schematic. Transistor substitutions may be made where those devices shown are not on hand.

The vco as well as a birdie marker system are combined on a single PC board. A schematic for this board appears in **fig. 1.** The mixer for the marker system is quite conventional and uses a dual-gate mosfet. This added feature is a practical necessity for most alignment operations and is complete except for the marker source. For precise measurements use a crystal oscillator to provide the marker signal; otherwise, an ordinary rf generator is quite adequate.

Transistor Q1 blanks the output during the retrace period and emitter-follower Q2 buffers the output from the vco. The output amplitude across a 50-ohm load is about 1.0 volt p-p maximum.

ramp generator

The improvements of which I spoke concerning an earlier design were made

When the output of U1A goes positive, the flip-flop consisting of Q3 and Q4 changes state so that Q3 turns on and Q4 turns off. The collector of Q4 is now positive and this voltage is fed back to Q2 via the voltage-follower U2A. When Q2 turns on, it rapidly discharges C1.

Comparator U2B senses zero and will flip when the voltage across C1 goes

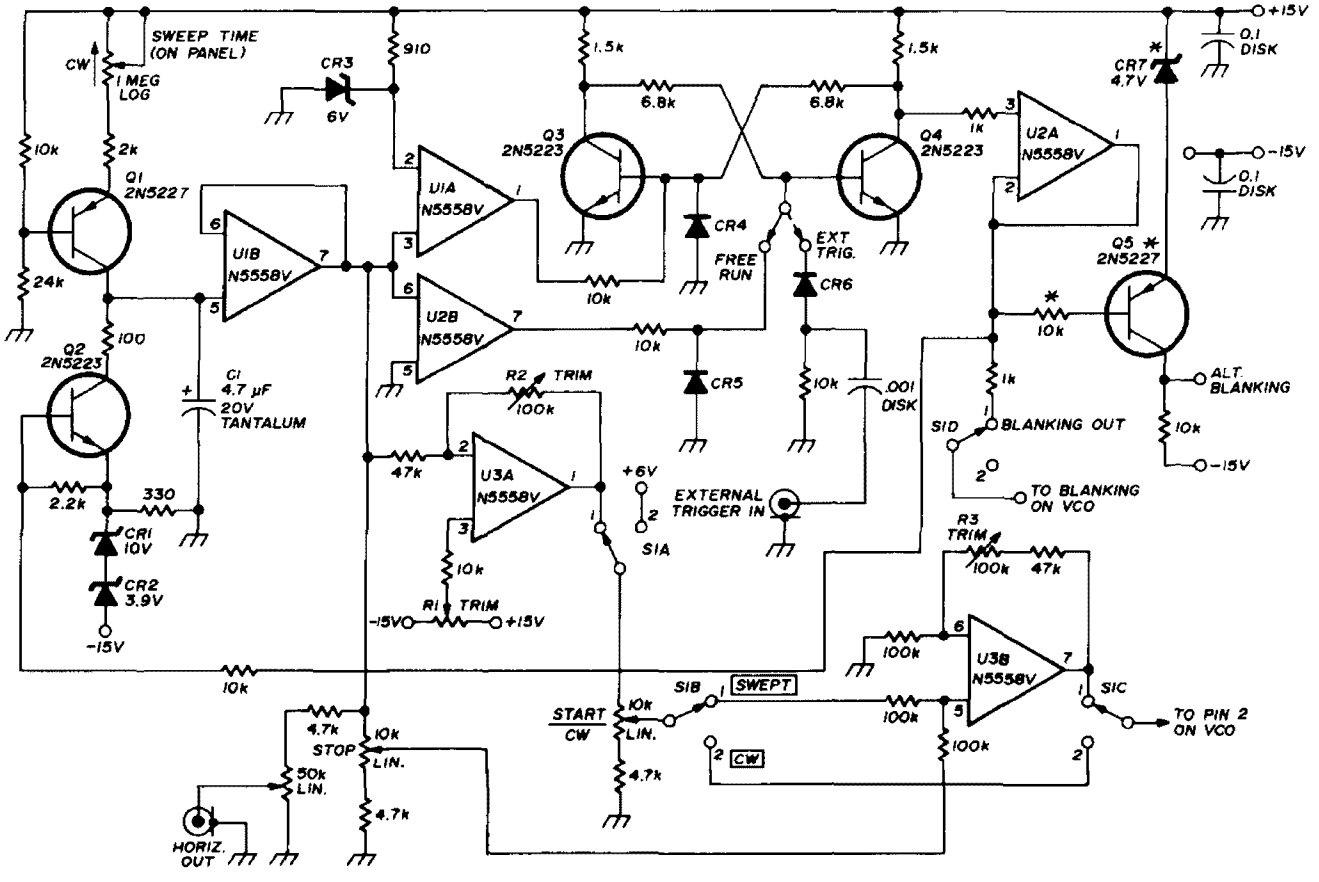


fig. 2. Ramp generator schematic. R1, R2 and R3 are Beckman 89PR100K 15-turn trimmers. All diodes are 1N914 or equivalent. Switch S1 is the mode switch. Components marked with an asterisk may be omitted as they are not required in this application.

mainly in the ramp generator circuit. This ramp generator is now my standard circuit for use in various types of sweeping instruments and is complete on its own PC board for that reason. A schematic of the ramp generator is shown in **fig. 2.**

A linear ramp is generated across capacitor C1 as it is charged by the constant-current source, Q1. This waveform, buffered by voltage-follower U1B, appears at the inputs to comparators U1A and U2B. U1A senses the positive peak and will flip when the level reaches a value just above that of zener diode CR3.

slightly below ground. This can happen since the emitter of Q2 is slightly negative. The output from U2B triggers the flip-flop which shuts off Q2 and allows C1 to start charging again as the cycle repeats.

If it is desired to trigger the ramp generator rather than have it free-running, a positive-going external pulse may be applied to the base of Q4 instead of the output from U2B. This is shown in the drawing and provided for on the PC board. If this feature is not to be included, the unnecessary components may

be omitted and a jumper installed from the base of Q4 to the output network of U2B.

In order to generate a voltage ramp whose end points can be set to any desired level, a positive-going ramp and a negative-going ramp are added together in a summing amplifier. As the amplitudes of the two input ramps are varied, the resultant ramp applied to the vco.

This circuit also provides two different blanking pulses. The one present at the output of U2A is a ground-based pulse which goes positive during the capacitor discharge period and is used for retrace blanking. The second blanking pulse is positive during the sweep period and goes

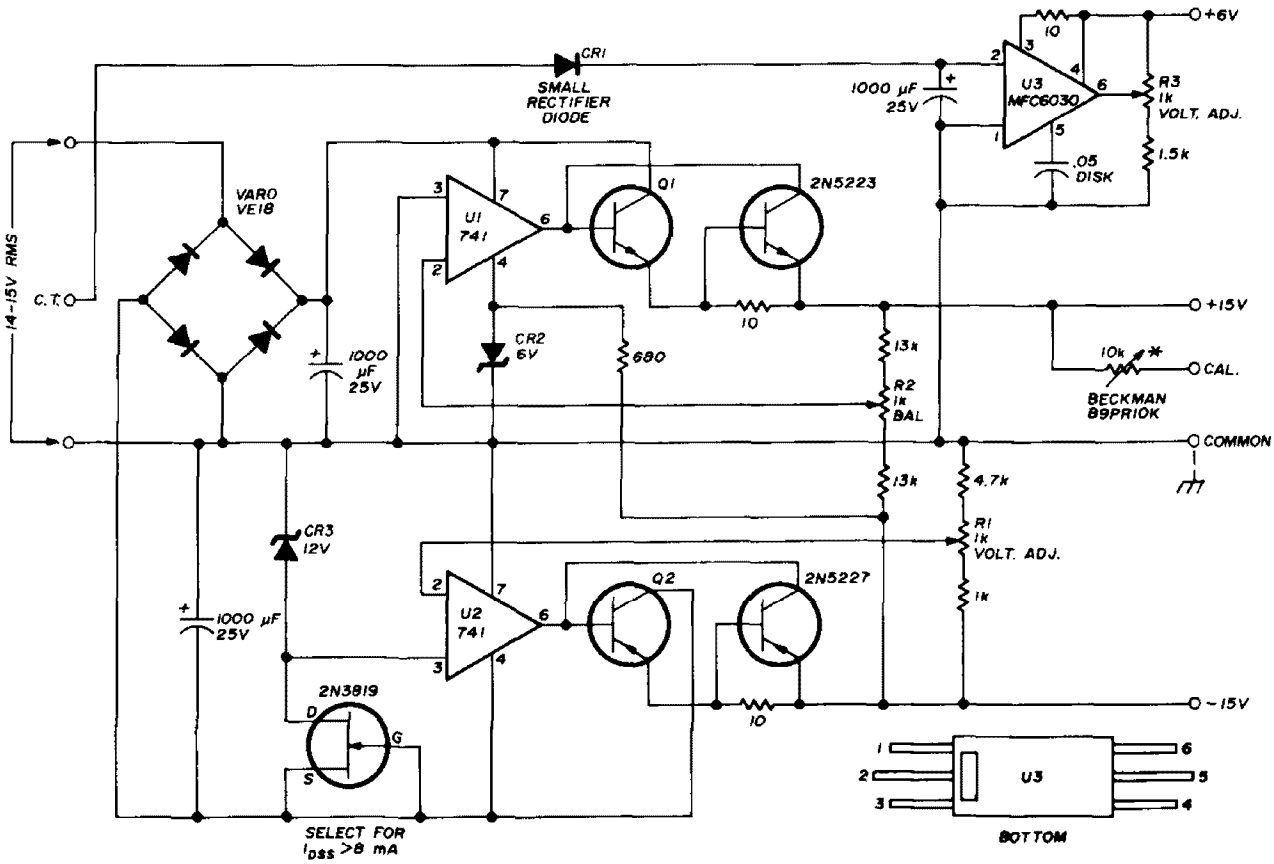


fig. 3. Power supply schematic. Transistors Q1 and Q2 should be silicon types with at least a 2-watt rating. U1 and U2 are 741 op-amps in T05 cans; remove pins 1, 5 and 8. U3 is a Motorola MFC6030. Trimmers R1, R2 and R3 are 1-turn vertical-mounting units such as the Wescom RV301. The 10k Beckman pot may be omitted as it is not needed in this application.

output will assume the desired characteristics. A more detailed explanation is given in my previous article and will not be repeated here.

The output from U1B is the basic positive-going ramp which is used as both the *stop* signal and the horizontal sweep for the scope. It is also applied to the input of amplifier U3A which inverts it and provides the negative-going ramp required for the *start* signal. Controlled amounts of these two waveforms are applied to summing amplifier U3B to negative for the discharge portion of the cycle. This latter pulse is not used here and, once again, the unnecessary parts may be omitted from the board assembly.

One possible innovation which may be of interest was incorporated into the original unit. By means of appropriate switching, a dc control voltage may be applied to the vco in place of the ramp and is varied in amplitude by means of the *start* pot. This allows CW operation of the generator much as any other rf source.

Although not provided here, the output could also be fm modulated by capacitance coupling an audio signal into the vco in place of the ramp. At the same time, CW tuning could be maintained to allow adjusting to any desired center frequency.

boards used in the original instrument. The vco board was laid out to accommodate a film-dielectric variable capacitor having three gangs, 330-pF each. With all sections in parallel, this unit provides a range of 20 to 990 pF.

Since it is unlikely that prospective

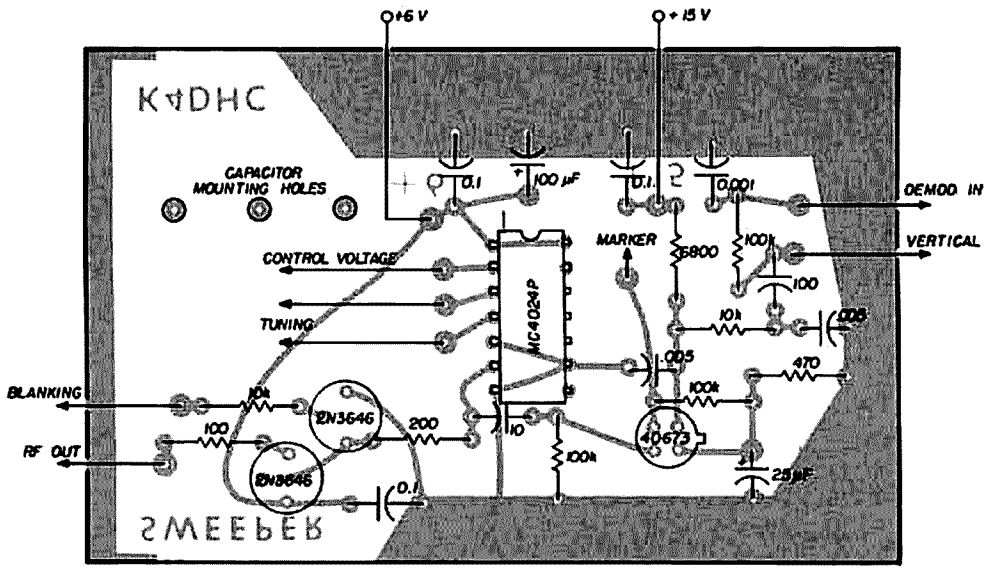


fig. 4. Component side of the vco board. Remove pins 10, 11 and 12 from the IC. Wide copper border on the left side is the common bus and should be grounded.

power supply

The power supply must furnish ±15.0 volts and +6.0 volts. Current requirements are quite moderate and no power transistors are needed in the regulators. Fig. 3 is a schematic for the supply I used. It is assembled on a PC board and is another of my standardized circuits.

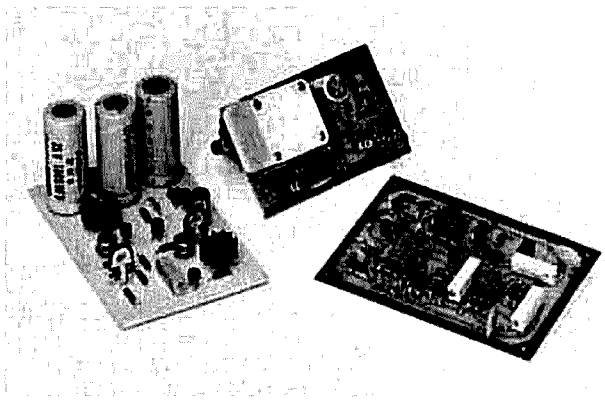
A tracking regulator is used in the ±15-volt section utilizing a pair of op-amps. Trimmer R1 will vary the voltage of both supplies simultaneously and trimmer R2 will vary only the +15-volt supply. Trimmer R2 is used to balance the two voltages and R1 is used to set the output level. The 6-volt section has its own regulator and trimmer.

Short-circuit protection is provided for all supplies. The 15-turn trimmer is not needed for the i-f sweep generator and may be omitted.

construction

Printed-circuit artwork and parts location drawings are shown for all three

builders will be able to find any of these variables (Mitsumi PVC-3R), I suggest using fixed capacitors and a range selector switch. Approximate values for a three-band assembly would be 820, 150 and 20 pF to cover the range from 100 kHz to 15 MHz. The use of compression trimmers is recommended for ease of alignment. The board can be fastened directly to the rear of the bandswitch in the space allocated for the variable capacitor.



The power supply (left), the vco (center) and the ramp generator (right) printed-circuit assemblies before installation.

If you decide to use a substitute variable capacitor and mount it external to the board, don't forget to insulate it from ground! This capacitor is internally connected to two emitters in a balanced circuit and neither end can be grounded.

All external connections to the PC boards are made by wire leads connected to the appropriate pads provided on the boards. There is one exception to this on the vco board. The 100-pF capacitor which couples the marker signal to the mixer, Q3, is itself used to complete the connection between the board and the front-panel connector.

The photos show the three PC assemblies before installation as well as overall views of the interior and exterior of the completed unit. The cabinet is a Ten-Tec MW-12 which I happened to have on hand, but it is really larger than needed.

alignment

The power supply voltages should be set to the proper levels before anything else. Balance the two 15-volt supplies by means of R2 and then set the level to 15.0 volts by adjusting R1.

The low voltage supply should be set to around 5.5 volts to start with as this may be adequate with some devices.

You may have wondered about the 4.7k resistors from the bottom end of each control pot to ground. These resistors overcome a 2-volt threshold in the vco control voltage response characteristic. Practically no frequency change occurs for the first 2 volts and a rather weird sweep would result if control was started from ground level.

When you are ready to start, set the mode switch to CW and the output frequency at the low end. With a scope or counter check the output frequency with the *start/cw* control at both minimum and maximum. If the tuning ratio is 5.5 to 1 or better, stop right there. If not, increase the supply voltage to around 5.8 volts and recheck to see if the ratio has increased. There will be a shift in output frequency due to the voltage change, but for now you are only interested in the

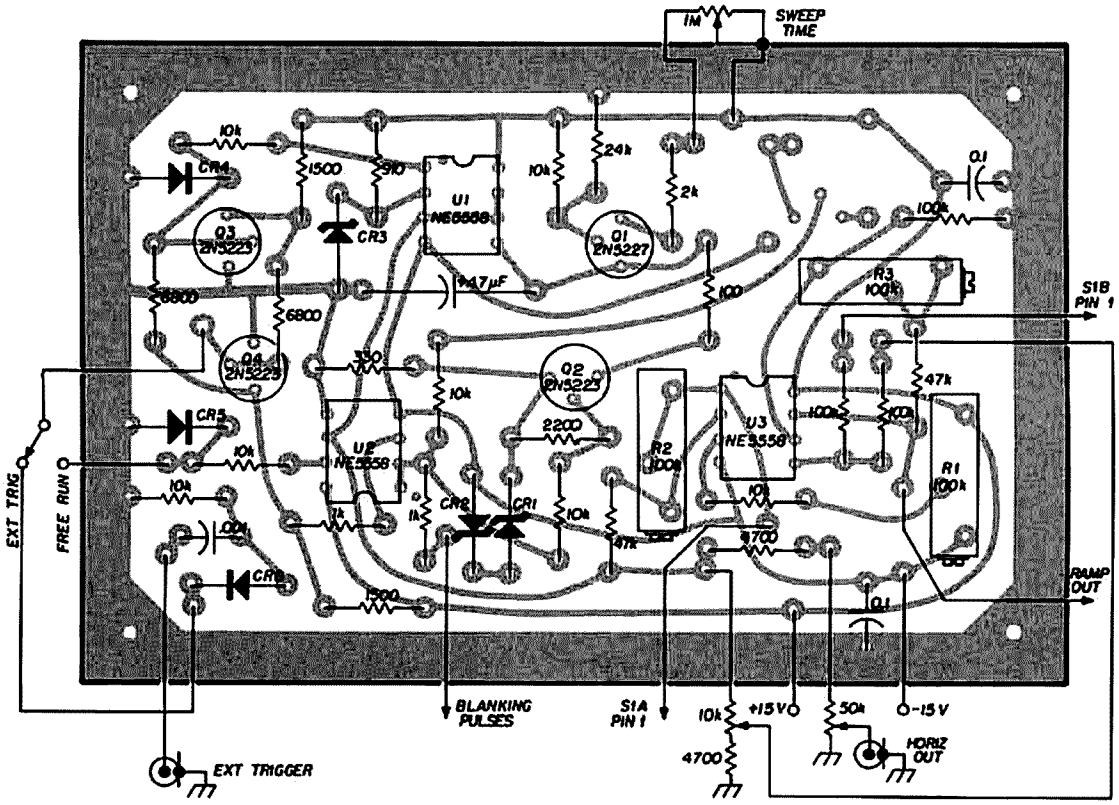


fig. 5. Component side of the ramp-generator board. The copper border is the common bus and should be grounded. If external triggering is not used, jumper points A and B on the board.

actual range afforded by the control voltage. Set the supply level for the best ratio but do not exceed 6.0 volts.

Once this phase is complete, return the *start/cw* control to minimum and proceed with marking the "base frequencies" on the panel. If you used trimmer capacitors and a bandswitch, now is the time to set them to the desired frequencies. These changes will not alter the tuning ratio afforded by the control voltage. In the case of a variable capacitor, mark appropriate spots on the dial to suit your convenience.

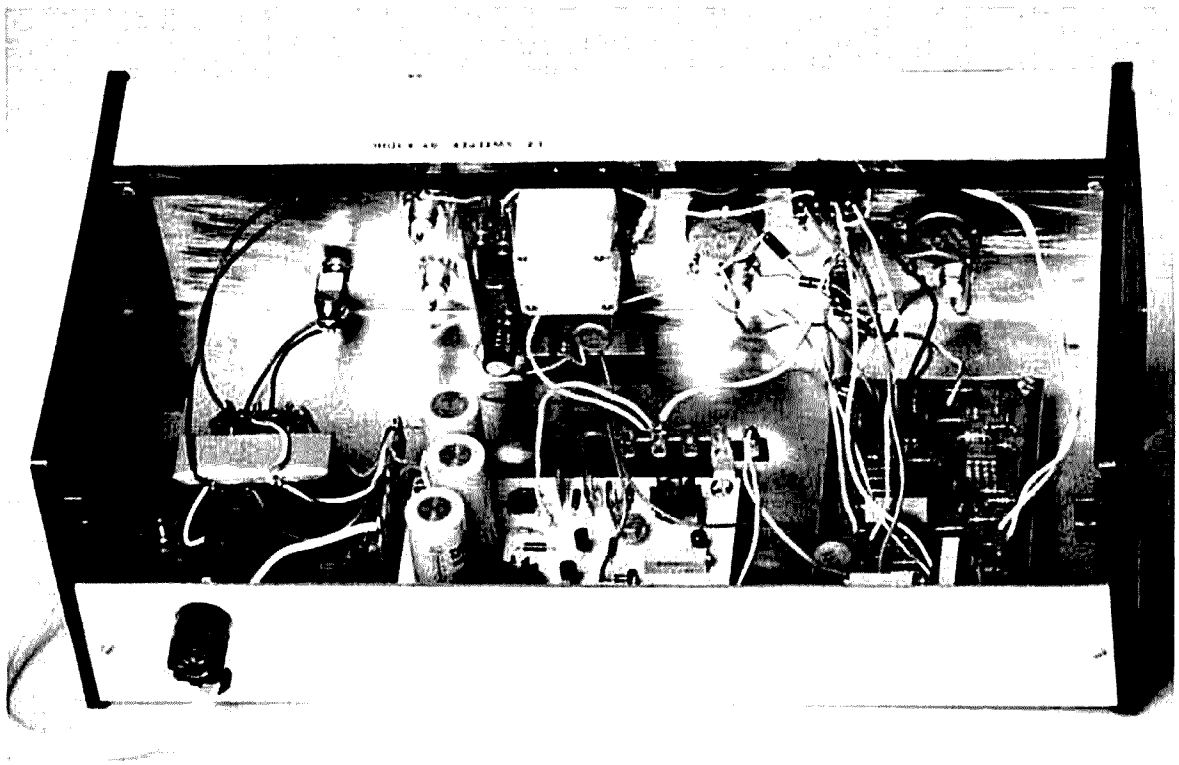
Next, the *start/cw* control should be marked off at points corresponding to multipliers of the base frequencies. These may be X2, X3 or whatever is suitable. Transfer these same points to the *stop* control and oscillator alignment is complete.

The adjustment proceedure for the ramp generator is next and will require the use of a dc oscilloscope. First, set the three 15-turn trimmers to mid-range, switch the mode switch to *swept* and turn the *sweep time* control to minimum resistance. Connect the scope probe to the top end of the *stop* pot. You should see a positive-going ramp of 6.0 volts amplitude and about 15 milliseconds duration with its base line at zero volts.

Switch the scope probe over to the high end of the *start/cw* control and observe a negative-going ramp. If all you see is a straight line, then it's likely the op-amp is latched up due to excess offset voltage. If the display is negative, start turning trimmer R1 in a counter-clockwise direction and reverse if the display is positive.

Once the ramp comes into view, adjust R1 so that the base line of the ramp is exactly zero. Next adjust R2 so that the amplitude of this ramp is the same as the first. You may have to readjust R1 since there is some interaction between these two trimmers. As soon as the two voltage ramps are set, connect the probe to a point on the mode switch where you can pick up the output ramp.

Turn the *start/cw* control to minimum



Interior view of the completed i-f sweep generator. The printed-circuit boards are mounted on metal spacers.

and the *stop* control to maximum. The result should be a positive-going ramp starting at approximately 2.0 volts and peaking at 6.0 volts. For tracking, adjust R3 so that the ramp peaks at the same voltage level as the maximum CW control voltage.

Turn the *start/cw* control towards maximum and observe the base line rising until, at maximum, there is a straight-line display. If any tilt still remains, it is due to summing error and may be corrected by readjusting R2 until the line straightens out.

Check for blanking pulses at the proper point on the mode switch and for the horizontal output ramp at the *width* control. Turn the *sweep time* control to maximum resistance. The sweep period should increase from the original 15 milliseconds to 5 or 6 seconds. That completes set-up and checkout of the ramp generator.

conclusion

Instruments such as the one I've described here can be extremely useful to the homebrew enthusiast and, even if constructed from all new parts, should cost only about fifty dollars.

Some portions of this device may be of interest even if the entire system is not. As I pointed out above, I have standardized the ramp generator and power supply for interchangeable use in several other designs. Incidentally, I normally use the full ramp capability of 12 volts in other applications and use a 12-volt zener for CR3. To maintain approximately the same sweep time range, a 2.2-μF capacitor is used at C1.

If anyone is interested in exchanging ideas regarding sweeping-type instruments, I would be pleased to hear from you. I will also try to help anyone having trouble locating any of the components or in duplicating the PC boards.

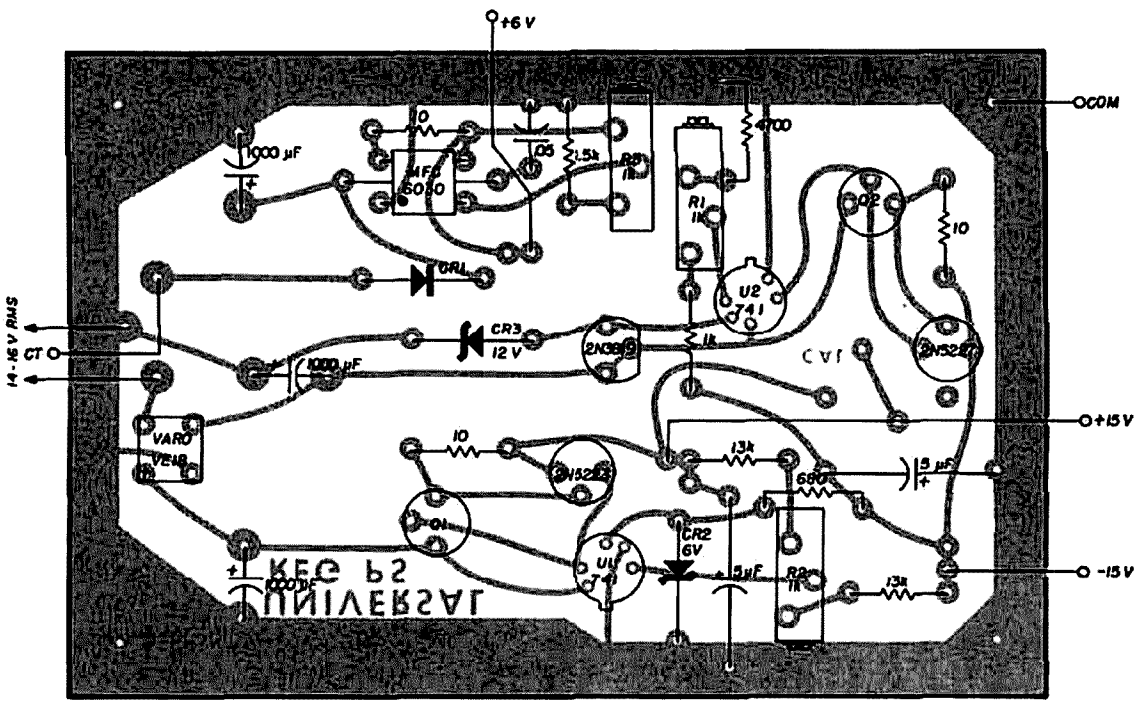


fig. 6. Component side of the power supply board. The copper border is the common bus and should be grounded.

reference

1. Ray Megirian, K4DHC, "Lab Type IF/RF Sweep Generator Using ICs," *73 Magazine*, November, 1972, page 226.

ham radio

rf speech processor for single sideband

Henry G. Elwell, Jr., W2MB, 392 Lafayette Avenue, Westwood, New Jersey 07675

High-performance rf speech processor for ssb offers 7 to 9 dB increase in effective audio output

Speech processing to improve communication capabilities has been discussed and developed for many years, and it covers both audio and radio frequency spectrums. Although audio clippers are simple to build and show effectiveness, they usually sound lousy. And, theoretically, they should.

Clipping produces square waves which are rich in harmonics. Thus, a 400-Hz tone includes 800 Hz, 1200 Hz, 1600 Hz, etc., up to the limit of the audio filter passband. Audio frequencies which do not produce harmonics in the audio passband, assuming a 3-kHz cutoff, are 1500 Hz and higher. The multiple audio harmonics produce the distortion in an audio-type clipper and give the high-pitched, tinny quality usually characteristic of such a device.

If the speech processor is used properly so that the transmitter presents a normal bandwidth signal, and if it truly enhances the communication capabilities of the station, it means little that your voice is high pitched and not recognizable to your friends — you are trying to improve your "get-through" ability.

However, some processors sound like "a one-man pile-up" and do little to improve communication effectiveness. An audio clipper I used improved my ability to get through the noise when signals were weak, but it sounded lousy. Studies and analyses indicate that processing at an rf level is the best way to go. During the 1972 ARRL DX Contest, a Comdel Speech Processor, which is based on rf

clipping, was borrowed from W2JKN. Its effectiveness proved superior to the audio clipper, and, in addition, voice quality was almost as good as the output from my un-processed SB100 ssb transmitter.

A number of articles have been written on the subject of rf clippers. The simple fact that sells me is that the filter following the rf clipper attenuates all harmonics of the carrier. If proper design precautions have been taken to build a high-fidelity audio system, the final detected audio signal should be good.

all harmonics of the fundamental frequency and restores the modulation envelope to its fundamental characteristics.

The product detector, which uses the same local oscillator as the balanced modulator, produces the processed audio. A lowpass audio filter provides desired audio shaping of the signal to the transmitter microphone input. Two sideband filters are required for proper rf processing — the second filter is the one already existing in the station ssb transmitter.

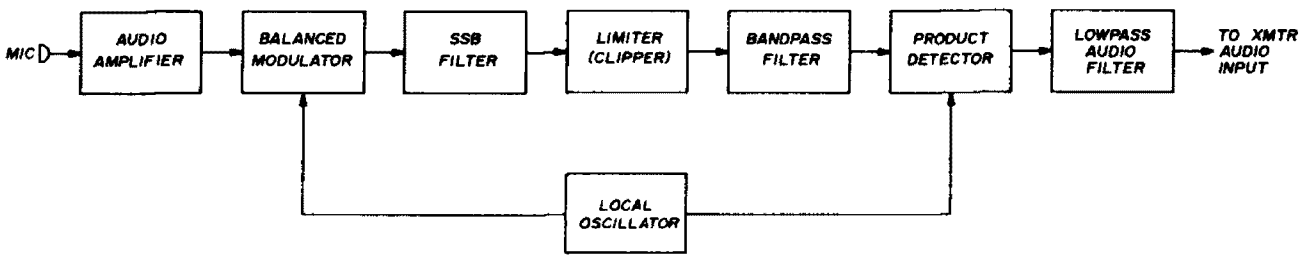


fig. 1. Block diagram of the rf speech processor.

My first attempt at rf clipping was in the intermediate frequency stages of the SB100, but it was unsuccessful. It's also awkward to modify a finished piece of equipment. The project lay dormant until my experience with the Comdel unit — then I had to have my own.

An rf clipper with audio in and audio out permits the construction of a black box which is put between the microphone and the audio input of the transmitter. It is actually a complete ssb transmitter-receiver, but its output is powerful (fig. 1).

operation

The audio amplifier amplifies the microphone signal to a level suitable to drive the balanced modulator. The balanced modulator, excited by a local oscillator, produces a double sideband signal with suppressed carrier. The ssb filter eliminates one of the sidebands, and the limiter provides the necessary gain to produce limiting or clipping at the carrier frequency. The bandpass filter attenuates

The nice part of the device shown in fig. 1 is that it can be built without disturbing the station transmitter. For all practical purposes each stage can be tested as you go along. However, you must build the stages in the following order to permit "as you go" testing: power supply, audio amplifier, balanced modulator and local oscillator, sideband filter, limiter and passband filter, product detector and, finally, the lowpass filter (see fig. 2). When built in this sequence, each item can be tested and certified as being correct before proceeding further. With this approach I was surprised at how little "fooling around" was required to make the device work properly when I connected it to my transmitter.

Many of us, including myself, get overwhelmed with the rapid strides in electronics these days, and we approach some of the new microcircuits with doubts in our minds, because it isn't a vacuum tube. I've taken a stand, however, that no more vacuum tubes will be used in any of my new equipment. Further-

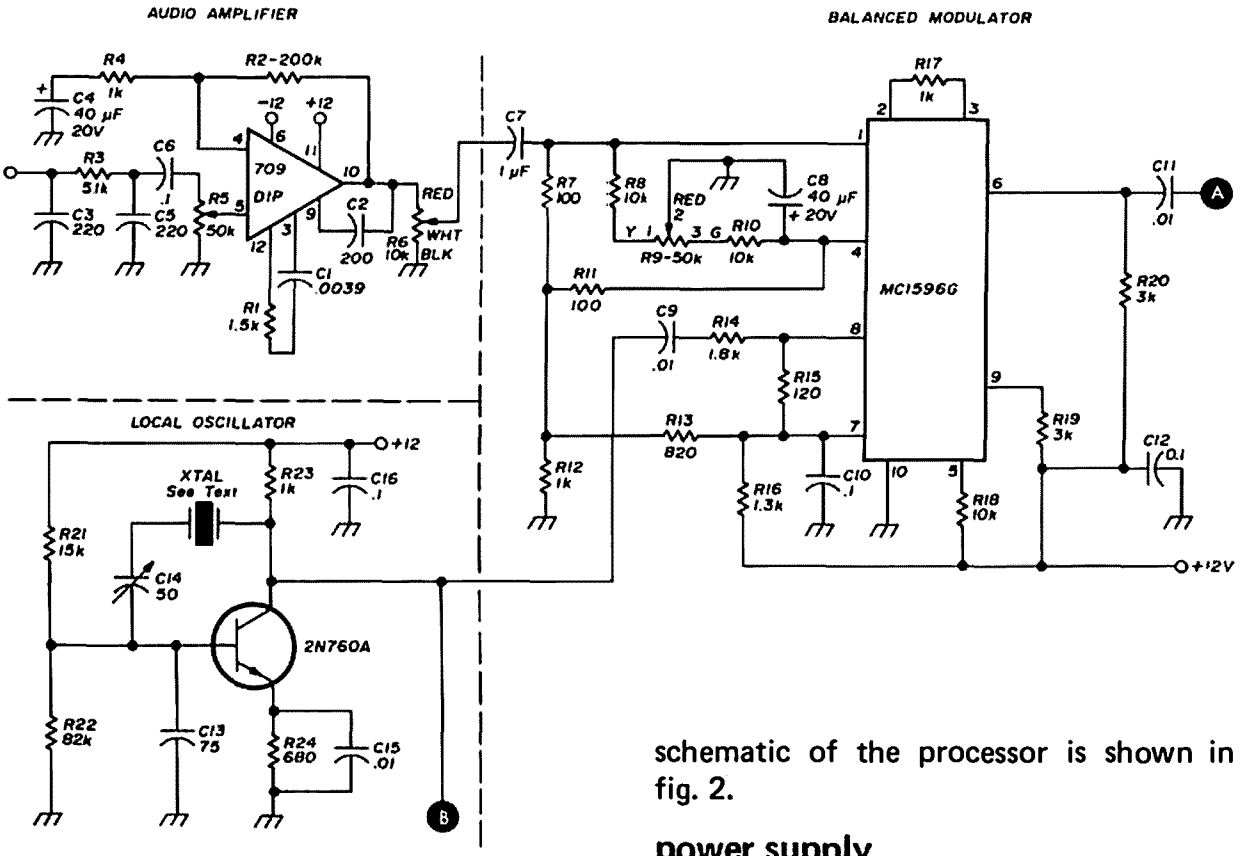
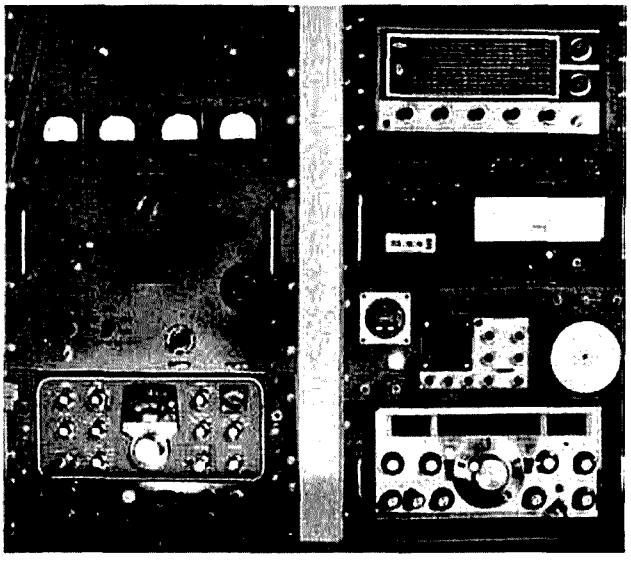


fig. 2. Circuit for the rf speech processor makes maximum use of integrated circuits. Power supply is shown in fig. 3.

more, if I can find a multifunction microcircuit to do the job, I will eliminate the transistors as well. Of course, this requires a lot of faith in the microcircuit application engineers. In building an rf speech processor the manufacturer's designs were followed by me explicitly, and the thing works great. A complete schematic of the processor is shown in fig. 2.



Station console used at W2MB.

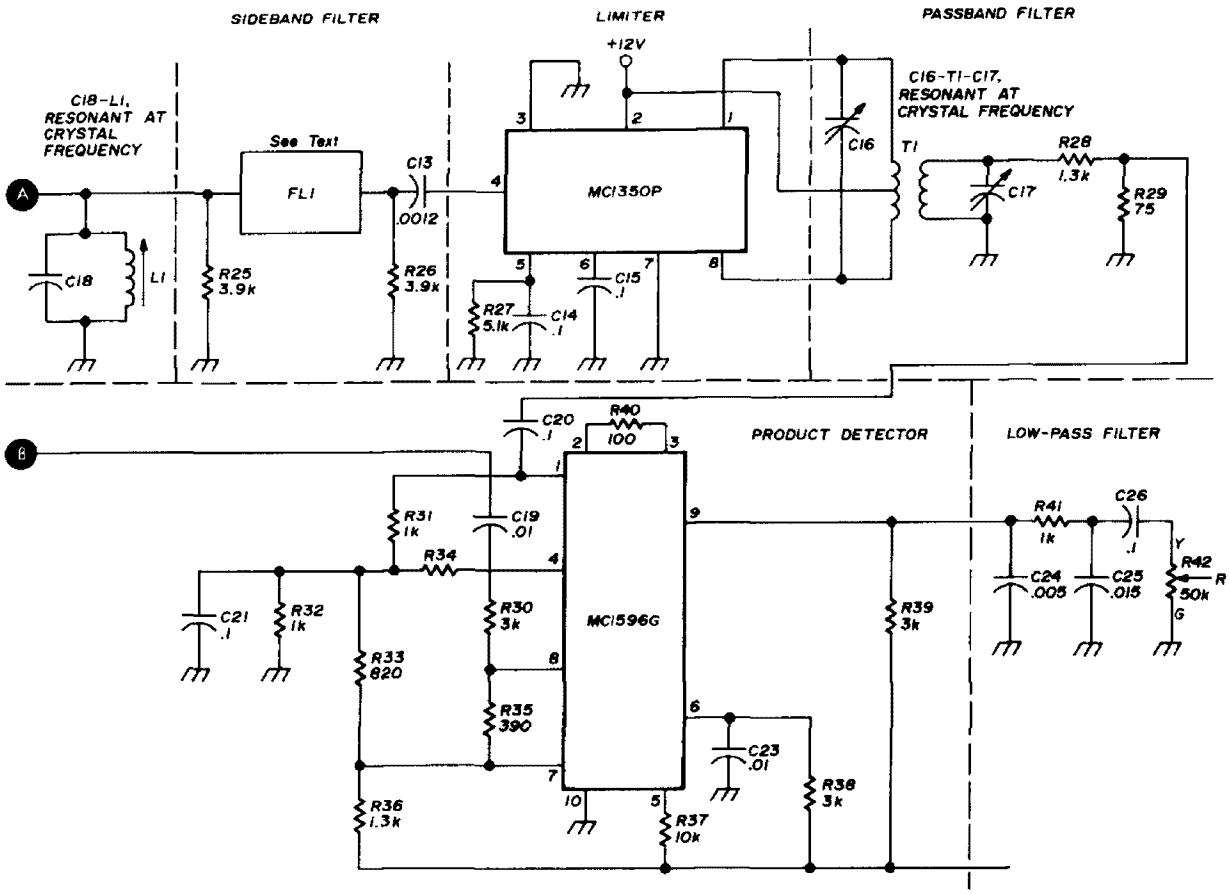
power supply

The processor circuitry requires both positive and a negative 12-volt supply. The negative 12 volts were required for the 709 audio amplifier; all other circuits require +12 volts only. To keep it simple, a transformer with a 24-volt center-tapped secondary was connected to a bridge rectifier and the center tap used as ground (see fig. 3).

To keep the output voltage within bounds, a 12-volt zener diode was used in each polarity leg. The selection of the series zener resistors was an experimental one, because the current requirements of the ICs were unknown at the time I built the power supply. Once the -12-volt load was established, it remained constant because the 709 IC was the only load. However, as additional loads were placed on the +12-volt line, the output level dropped below 12 volts because of the drop across the series zener resistor. The value of that resistor was then lowered to permit the zener diode to regulate the output voltage. The ripple on the positive voltage is higher than the negative output, and a larger filtering capacitor is used.

audio amplifier

The design philosophy of the audio



oscillator was to obtain as flat a frequency response and low distortion as possible, relying on the sideband filter to provide the desired audio response. Just any old IC op-amp is not suitable as some of them have poor frequency response. The 709 was used because it has good response and is readily available. The circuit, as shown, is flat, within 2 dB, from 30 Hz to at least 10 kHz, with a gain of about 200. Maximum undistorted output is 7.5 volts.

The potentiometer, R5, is used for setting the clipping level, whereas potentiometer R6 is used to set a maximum permissible level to the balanced modulator. An rf suppression network, consisting of C3, R3 and C5, was included to minimize rf feedback problems. However, it was not entirely effective as will be discussed later.

local oscillator

The choice of the carrier frequency is your own. I had two crystal filters on hand; the original SB100 filter, 3.395 MHz, removed *from* the rig when I installed the Heath 400-Hz filter, and a Hermes model 2215KZ filter, centered on 2.215 MHz. The Heath filter has a nominal shape factor of 2.1 while the shape factor of the Hermes is 3.5. Since the 2.1 shape factor may be needed on some other, more critical, project, the Hermes filter was used in the speech processor. It makes little difference what frequency your filter is as long as it is not too high. The limiter stage should have as large a spectrum above its operating frequency as

table 1. Voltage readings at the pins of the μA796 balanced modulator.

pin number	dc reading (volts)
1	3.6
2	2.9
3	2.9
4	3.6
5	1.2
6	8.7
7	6.8
8	6.8
9	8.7
10	0

possible to permit it to generate the many harmonics that result in effective clipping action. Thus, the lower the better seems appropriate.

The Comdel speech processor uses an oscillator frequency below 500 kHz, balanced-modulator circuit I used was described in an industrial electronics magazine, *EDN/EEE.*[1] The local-oscillator signal is recommended as being 60 mV. Adjustment of the local-oscillator signal is accomplished by varying R14 and R15 to

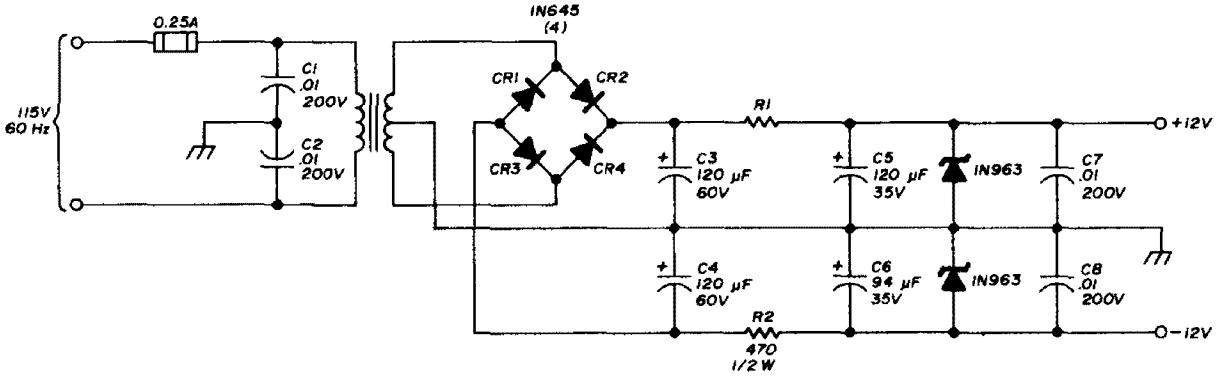


fig. 3. Power supply for the rf speech processor. For dropping resistor R1, start with 470 ohms (2 watts) and add 390-ohm resistors in parallel as more stages are added to the processor (see text).

which permits the use of an LC sideband filter. Some readers may wish to explore that area. However, since I had the crystal filter on hand I used it.

I used a crystal oscillator of conventional design, but the use of a trimmer at C14 is essential to properly align the crystal frequency in the crystal filter passband. A total variation of about 1 kHz was possible with the trimmer shown in fig. 2. The oscillator has an output at the collector of 2.4 volts.

balanced modulator

A number of articles have been written using the Motorola MC15966 balanced modulator.* The IC I actually used, however, was the equivalent Fairchild µA796, because of its availability. The

table 2. Dc operating voltages of the MC1350P IC under quiescent conditions.

pin number	dc reading (volts)
1	9.0
2	12.3
3	0
4	4.0
5	4.2
6	4.0
7	0
8	9.0

prevent excessive loading of the oscillator while, at the same time, keeping R15 as low as possible; 50 ohms is recommended. It was a compromise, therefore, to end up with a local oscillator input of 85 to 90 mV with R15 at 120 ohms.

A dc check was run using three different µA796 ICs and they all have the same readings as shown in table 1. Measurements were made with a Heath Model V-7A-volt ohmmeter.

The recommended modulating signal voltage is 300 mV and R6 is used to set the proper level. The setting of R6 depends upon the microphone being used, and it is necessary to experimentally determine the microphone output. Then, with an audio signal input at the maximum expected level and with R5 at one-half range, R6 is adjusted to produce a µA796 pin 1 level of 300 mV. Output at pin 6 with C18-L1/ peaked is about 200 mV.

The null balancing potentiometer R9 with zero audio signal is adjusted to produce minimum output voltage. The best I could obtain was 30-dB attenuation.

*The National LM596, Signetics NE596 and Fairchild µA796 are equivalents.

sideband filter

After the sideband filter is wired in, and output is obtained from it, the frequency of the local oscillator must be adjusted for the desired passband. The setup shown in fig. 4 is useful.

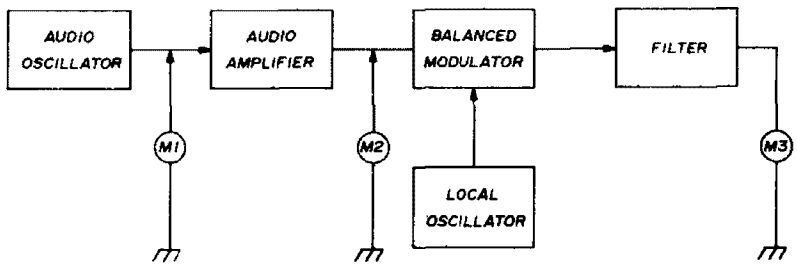


fig. 4. Test setup for setting the frequency of the local oscillator for the desired audio passband. Meters M1, M2 and M3 are voltmeters.

The voltmeter, M2, is not really necessary, but it provides information on the response of the audio stage. A frequency response curve from about 50 Hz to 4000 Hz should be run and plotted in dB vs frequency. The local oscillator capacitor, C14, should be adjusted to move the audio passband up or down, depending on the initial frequency plot. My initial attempt put the lower frequency roll-off at 600 Hz; this is much too high. Since my voice is normally high pitched, the oscillator frequency was adjusted so that the 6-dB down points occurred at 200 Hz and 2700 Hz.

iimiter

I originally decided to use a μA709 IC op-amp for the limiting stage. However, after considerable attempts at high-frequency compensation with poor results, the idea was abandoned. The use of the μA709 op-amp was carried through, however, until the complete processor was finished and put on the air. It sounded raucous and was quickly removed.

An LM373 was tried as a combination clipper-filter-product detector stage, but it was also unsuccessful. The LM373 would be ideal if the input to the product detector was available for monitoring or filtering.

Finally, a limiter using an MC1350P was built. The MC1350 is a $1.15 item designated as an i-f amplifier. It has a typical power gain of 48 dB at 58 MHz and was put to work in a circuit recommended in the Circuit Specialists catalogue.

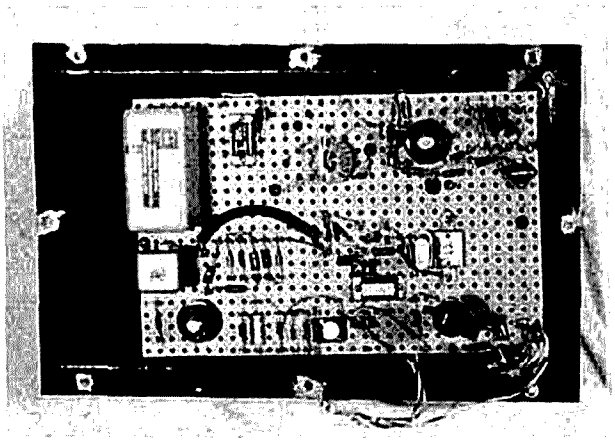
The device has one feature which I first felt to be a disadvantage – it requires a center-tapped primary for the output transformer. I felt it was desirable, for simplicity, to use a single-parallel tuned circuit for the passband filter as was done on the output of the balanced modulator. However, the MC1350 output requirement made it necessary to use a double-tuned transformer with much superior passband characteristics. The output from the filter is an exceptionally good sine wave, even at maximum limiting, indicating excellent harmonic attenuation. A conventional shielded, slug-tuned coil form was originally used but was replaced by a small toroid for physical size reasons only.

Tuning of the primary and secondary of the toroid coil is accomplished with trimmer capacitors. As oscillation occurred with the secondary open; a 1.3k ohm resistor across the secondary stopped the oscillation. Maximum voltage across the

table 3. Dc operating levels of the μA796 product detector under quiescent conditions.

pin number	dc reading (volts)
1	3.8
2	3.0
3	3.0
4	3.8
5	1.3
6	9.0
7	7.2
8	7.2
9	9.0
10	0

secondary is 2.4 volts. Limiting of this stage occurs at 80-mV audio voltage into the balanced modulator, but no distortion appears across the secondary winding of the transformer at 800-mV input to the balanced modulator. Table 2 indicates the dc voltages to be expected under quiescent conditions.



Speech processor is built on a section of perforated board. Crystal filter is in upper left-hand corner with MC1350 limiter to the right. MC1596 product detector is in upper right, just above crystal-controlled local oscillator. MC1596 balanced modulator is in lower right-hand corner.

product detector

A μA796 IC is also used as a product detector to convert the rf signal back to audio. Again, the recommendations of the application engineers were followed and the circuit worked fine. Table 3 indicates the dc voltages to be expected under normal quiescent conditions.

The designers' recommendations regarding ssb level to the product detector are to use up to 100 mV rms. Since the output from the limiter stage was excessive the resistive loading on the secondary of the rf transformer was made into a voltage divider to supply a maximum of 100 mV to the μA796. The carrier level was set at about 180 mV by means of the voltage divider feeding pin 8 of the IC.

lowpass filter

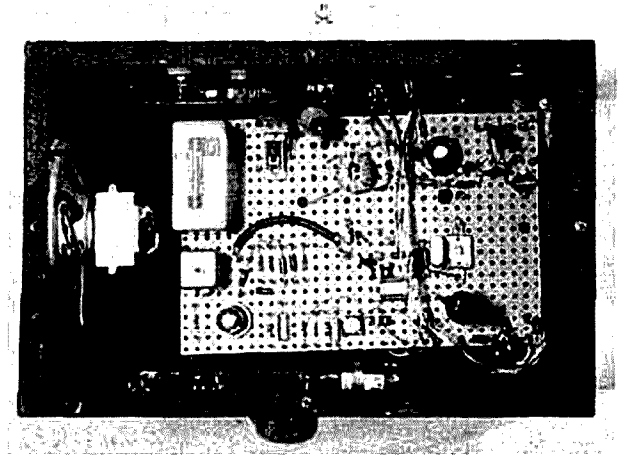
The lowpass filter eliminates all remaining rf energy and provides a little audio shaping. I used a 0.015-μF output capacitor to provide better balance between my normally high-pitched voice and the permissible frequency response of the system. The output potentiometer, R42, was adjusted so that the processor output voltage was the same as that delivered by the microphone directly. Thus, I can switch from normal to processor with no adjustment to the transmitter microphone gain control.

dummy load tests

Tests with the SB100 ssb exciter into a dummy load indicated that the audio with the new processor in the line sounded much like the un-processed SB100 signal, with one difference — the audio was very loud and full. To see what was happening, different limiting values were cranked in and measurements recorded from the SB100. The results are shown in table 4.

How was the dB limiting cranked in? W2LL and I discussed the meaning of "a given dB of limiting," and made the following conclusions. First, increase the audio input signal until limiting of the output occurs, call it A. Then, the ratio of a further increase in input audio to the A input signal, call it B, expressed in dB is the amount of limiting in dB.

This was done by measuring the audio voltage at pin 1 of the balanced modu-



Speech processor circuit board is mounted in an old portable receiver chassis (speaker on left is not used).

lator, while monitoring the audio level at the output of the processor. With an input signal equal to maximum mike level, increase R5 from zero output until a limiting output is obtained at the processor output. Just juggling R5 while watching the output meter will give a suitable answer. Call this measured input voltage to the balanced modulator **A**. If you double the **A** voltage by increasing R5 you have produced 6-dB of limiting. Of course, the processor output does not change.

If you increase the **A** voltage ten times, you have 20 dB of limiting or clipping. That is the way I did it and, having read no other description elsewhere, the procedure is assumed to be correct.

The tests of **table** 4 were accomplished by talking into the microphone and noting the peak of the two meter indications; plate current and relative power. It was necessary to use a repeatable sound, and the word "three" was found to be the best of the usual "one, two, three, four, hello test," variety. The audio gain of the transmitter was adjusted so that grid current of the output stage would barely flicker. The adjustment was tricky with the microphone alone, but, with the processor, there was no problem. Once limiting occurs, no additional grid drive is experienced, regardless of how loud you talk. However, the test showed that increased limiting did produce higher plate currents and relative outputs.

on the air tests

W2KXD, who is my strongest critic in the speech processing area, said that my signal was 1000 times better (whatever that means) than my audio clippers or my earlier test with the 709 limiter and was acceptable for on-the-air use. Brief tests were made on 75 meters and without the processor my signal was S9 +29 to +34 dB. With the speech processor in the line my signal was S9 +38 to +41 dB. This is an improvement of 7 to 9 dB in signal strength.

table 4. Performance of the speech processor with different limiting levels.

limiting (dB)	plate current (mA)	relative power
none	100	4.0
8	140	5.5
14	160	6.5
17.5	170	7.5

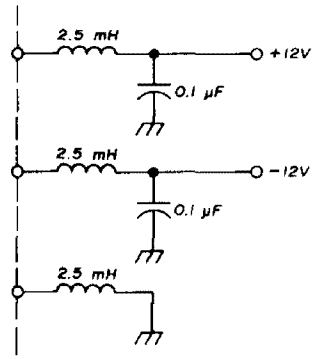


fig. 5. Filtering added to the power supply leads to prevent rf feedback within the speech processor.

Next, I went to 20 meters and got into a pileup for a UK2. He came back to me and we exchanged reports and names; before we could sign a UK3 called me so we moved up 5 kHz and talked for awhile.

Then I went to 15 meters and calamity hit — rf feedback galore. Ten meters and even 40 gave me problems. The processor was in a metal box, the power supply was in an external metal box and the interconnecting wires were plain hookup wire. The solution was to put a 2.5-mH rf choke between a feedthrough on the processor box and a 0.1-μF ceramic capacitor to ground for the +12 and -12 volt lines and the ground lead as shown in fig. 5. This completely cured the rf feedback. The microphone input cable and the cable from the processor are shielded.

ham radio

reference

1. Roy Hejhall, "For High-Frequency Communications Use Balanced Modulators," *EDN/EEE*, February 15, 1972.

coax dehumidifier

Here's a drier for hard-line coax that uses no motors, no pumps, no electricity, no gas, and practically no money

William L. Smith, K4RJ, Franklin, North Carolina 28734

It looks like RG-8/U coax is on the way out, and not a day too soon. Repeater operators, meteor chasers, aurora gazers, moonbouncers, microwave nuts, and even dc-band DXers are making more use of low-loss lines which use metal-tube outer conductors in place of that old braid.

There are two kinds of hard-lines in general use. In one type the cable is filled with foamed insulating material, which excludes air and moisture, and supports the inner conductor. In the second type, usually called "air dielectric" the inner conductor is supported by a minimum of insulating material, the rest of the space being filled with dry air or gas. This is the kind which has the lowest losses at all frequencies.

The big problem with air-dielectric cable is how to keep it dry inside. Commercial installations use tanks of dry nitrogen or other gas under pressure, air pumps equipped with dessicators, or mechanical dehumidifiers, and sometimes even include heaters for drying the dessicant. Most such installations cost between $100 and $1000; amateurs need a more simple and economical solution to the problem.

The cable I use at K4RJ (and formerly at W3GKP) is a 40-foot length of 1-5/8-inch Styroflex. The input to the cable is about 275 watts at 2304 MHz. The power is monitored with a directional coupler which gives a full-scale reflected-power reading (and automatic transmitter

switch-off) at approximately 1.2:1 standing wave ratio. The loss in the line is about 2 watts per linear foot.

observations

Four years of operating produced the following observations:

1. If the cable is sealed at both ends while filled with ordinary air containing some moisture, the swr is excessive. If a tuner is inserted to lower the swr seen by the transmitter, the swr readings become unstable at full power. Each adjustment of the tuner to lower the swr results in another shift in impedance, the condition continuing for several hours, possibly without end. It seems that the power dissipated in the line is moving the moisture around from one place to another.

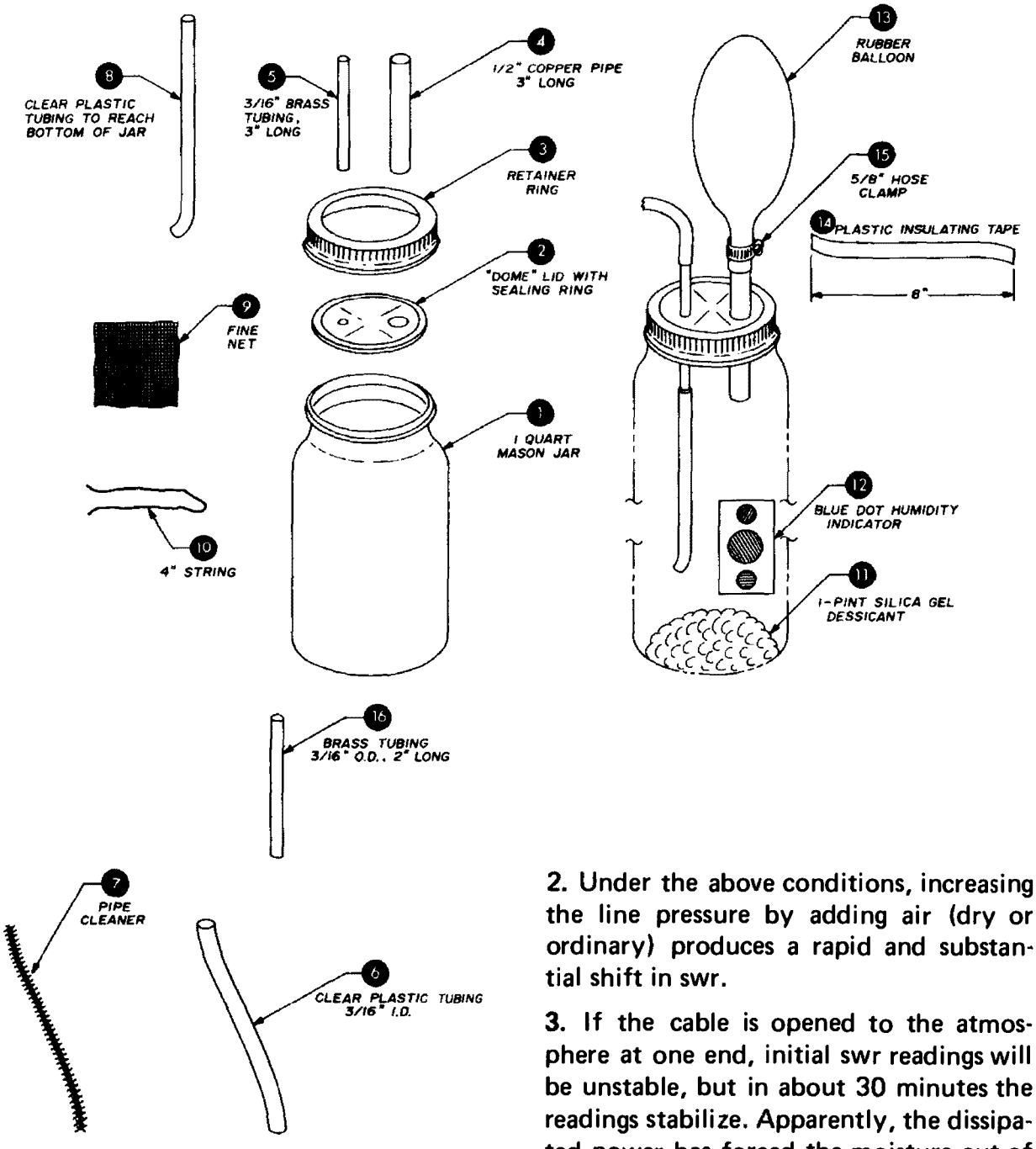


fig. 1. Construction details of the low-cost hard-line dehumidifier. See table 1 for complete list of parts.

2. Under the above conditions, increasing the line pressure by adding air (dry or ordinary) produces a rapid and substantial shift in swr.

3. If the cable is opened to the atmosphere at one end, initial swr readings will be unstable, but in about 30 minutes the readings stabilize. Apparently, the dissipated power has forced the moisture out of the cable. If the cable is left unsealed the whole job must be repeated the next day.

4. Opening the cable at both ends and flushing it with dry air (air from a small compressor pumped through a can of dessicant and into the cable) will result in stable readings. However, this air flushing takes longer than the rf flushing described in 3 above.

5. If one end of the cable is sealed, and the other end is attached by a short length of tubing to a closed can of dessicant, the cable can be flushed by rf in about 30 minutes, as in 3. Thereafter, the swr will remain low for a week with no further operation of the transmitter.

objectives

The objectives for a home-made drying system would be to keep the line under a slight positive pressure (to insure there is no leakage from outside in), and should include a pressure indicator and an over-pressure release. It should also include a drying agent and a humidity indicator.

Maintenance, if required, should be simple and easy. Periodic operation of any of the apparatus should not be needed. Furthermore, the drying system should be safe, and both initial cost and operating expense should be minimized.

My system meets these objectives. One end of the cable is sealed; the other end is connected by tubing to a container of dessicant. Normal changes in temperature cause the cable to "breathe" through the dessicant. Air "exhaled" by the cable passes into an ordinary balloon, which serves as a constant-pressure variable-volume indicator, and safety valve. The dessicant is held in a glass container, through which the humidity indicator can be seen clearly.

construction

All the parts for the dehumidifier are shown in fig. 1 and described in table 1. First, punch, drill or ream two holes in the dome lid to pass the 3/16-inch brass tubing and 1/2-inch copper pipe with a snug fit. Tin the edges of the holes with solder, insert the two pieces of tubing, leaving about a third of the length on the lower side of the lid, and solder them in place. Be careful when soldering not to damage the rubber seal around the edge of the lid.

Now, attach the clear plastic tubing to the 3/16-inch brass tubing, and assemble the lid to the Mason jar with the retaining ring. Submerge the jar in water, and while holding your thumb over the open end of the copper pipe, blow into the free end of the plastic tubing and look for air bubbles. If it leaks, re-solder the joints. If it's airtight, separate all the parts and dry them thoroughly.

table 1. Parts list for the coax dehumidifier.

1. 1 quart Mason jar
2. 1 Dome lid for Mason jar
3. 1 retaining ring for holding Dome lid on jar
4. ½" copper pipe, 3" long
5. 3/16" OD brass tubing, 3" long
6. 3/16" ID (5/16" OD) clear plastic tubing, length to reach from dehumidifier to coaxial cable
7. pipe cleaner
8. 3/16" ID (5/16" OD) clear plastic tubing, length to reach bottom of jar
9. 2" square nylon stocking material or fine net
10. 4" string
11. 1 pint silica gel dessicant
12. Blue-Spot humidity indicator
13. rubber balloon, size and color optional
14. 8" length plastic insulating tape
15. 5/8" stainless hose clamp
16. 3/16" OD brass tubing, 2" long

Wrap the fine net (an old nylon stocking works fine) over one end of the short section of clear plastic tubing (part 8 in fig. 1) and tie it in place with the piece of string. Push the other end of the plastic tubing over the lower end of the 3/16-inch brass tubing which is soldered to the jar lid.

If the silica gel has been exposed to the air, dry it according to the instruc-

tions on the container (or spread it on a flat pan and bake it in an oven at 250° F for five hours). When the dessicant is completely dry, place the Blue-Spot humidity indicator inside the Mason jar so the spots are visible from the outside and pour the silica gel into the jar. Reassemble the lid and retainer ring on the jar and work the plastic tubing down through the silica gel until it's near the bottom of the jar.

Place the rubber balloon over the top of the 1/2-inch copper tubing, wrap it tightly with the plastic insulating tape, and secure it with the 5/8-inch hose clamp. Slip one end of the long section of plastic tubing (part 6, fig. 1) over the end of the 3/16-inch brass tubing. Place the pipe cleaner (a secondary dirt filter) in the free end of the plastic tubing.

Arrange the air fitting on the coaxial line to receive the free end of the plastic tubing. If necessary, drill or ream the air fitting to accept a 2-inch length of 3/16-inch brass tubing (part 16) and solder the tubing to the coaxial line. The plastic tubing can be attached to this brass tubing.

Now, blow into the free end of the plastic tubing until the rubber balloon is suitably inflated. Quickly push the free end of the plastic tubing over the brass tubing soldered to the coax line. This completes the dehumidifier system.

Obviously, there is nothing magic about the dimensions I used, but a few remarks may be in order. The flexible plastic tubing should be a snug push fit on the brass tubing so that clamps are not needed. The shorter the length of the tubing to the coaxial line, the better the system will work.

If the balloon is a snug fit on the copper tubing, it may be held tight enough just by the tape. In one case I found the balloon fit *only* loosely, so the hose clamp had to be used.

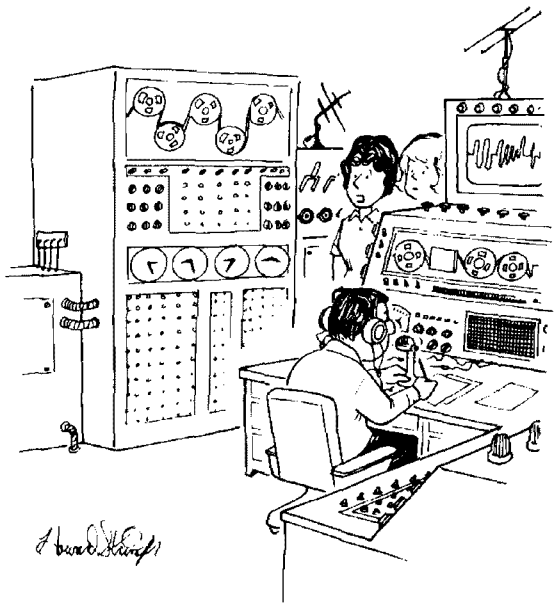
If the system is inflated by lung power, condensed moisture from the breath may be seen inside the plastic tubing, especially in cold weather. However, this will dissipate in a few days. It can be avoided by using a tire pump or football pump to pressurize the system. Initial drying of the air can be accelerated by running the line at full power for an hour or so.

Don't expect the size of the balloon to change dramatically unless you have a long, fat coaxial line. The actual change will depend on the temperature variations. Black-jacketed line exposed to the sun will heat up much more than line buried in the earth.

On a recent trip I described this system to several amateurs and one asked, "What happens when the balloon breaks?" I thought this was a joke until I returned home and found the balloon split and deflated (it was an old one). This is not a crisis because the cable continues to breathe through the dessicant, which traps any moisture which might otherwise enter. My system sat "open" like this for four days without any noticeable change in the dryness of the dessicant or the swr of the cable.

Now that we have gotten rid of that RG-8 – what about the PL-259?

ham radio



"It started out as a simple listening post. But that's how Arnold is. . . When he gets involved in something, he goes all out."

one-crystal frequency-synthesizer for two-meter fm

Design aspects of a frequency synthesizer for receiver and transmitter frequency control on two meters

Jack Cox, WØKMV, 13525 West 7th Drive, Golden, Colorado 80401

Integrated circuits have made it possible to build simple frequency synthesizers that heretofore had been complex to design and construct. As an added bonus, 7400-series TTL ICs have made even a large synthesizer inexpensive. Several articles describing synthesizers have appeared, but most of them are designed for a specific transmitter or receiver or transceiver. Modifying them to fit other frequency-multiplication and i-f schemes has called for a major redesign of the circuit. In addition to these problems, some frequency synthesizers have an extremely long lock-up time, making them unusable for transceive operation.

The heart of a digital synthesizer is the phase locked loop (PLL). Basically, the PLL is a circuit that compares the phase of two input signals and produces a dc voltage whose amplitude is relative to the difference in phase of the two signals. This dc voltage, after filtering, is used to change the frequency of a voltage-controlled oscillator (vco). When the vco frequency and phase reach a value equal to the reference frequency applied to the PLL, the dc voltage stops changing and the loop is said to be locked. A basic phase-locked loop is shown in fig. **1**. From this circuit will be developed a frequency synthesizer that is flexible and can be adapted to almost any transceiver presently in use.

Although this article is not intended as a construction article, all of the circuits shown have been breadboarded and work. A synthesizer using these circuits is now under construction. This article is intended primarily to stimulate thought in the design of synthesizers in order to obtain the best possible circuitry.

basic phase-locked loop

Referring to fig. **1**, it can be seen that the output frequency of the vco and the reference frequency must be the same to arrive at a locked condition. Although

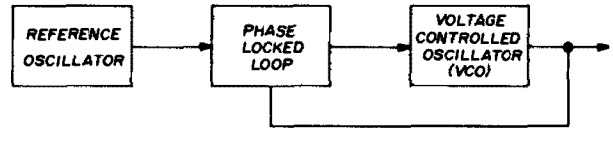


fig. 1. Basic phase-locked loop.

this is not practical for our purposes, it does bring to mind the possibility of using the PLL as a stable frequency reference by using a conditioned signal source.

Note also that the vco output frequency will be just as stable as the reference frequency input to the PLL. In any frequency synthesizer, stability is dependent upon the reference frequency and should be given serious thought. Since the circuit of fig. 1 is *unusable in a* channelized system, it is expanded to include a counter or programmable divider, as shown in fig. 2.

By including the counter in the feedback loop, you can now produce N number of output frequencies with only one reference frequency input. This is achieved by programming the counter to always have a frequency output equal to the reference frequency, regardless of the vco output frequency.

For example, assume the reference is 1 MHz and the desired output frequency is 7 MHz. The programmable counter, in this case a down-counter, is preloaded with the number 7. The counter now counts down from 7 to 0 and reloads, and the count recycles again. The counter produces one output pulse for every seven input pulses. This is equal to 1 MHz output for 7 MHz input and the loop is locked, producing a stable output at 7 MHz. Lock is achieved by the PLL producing a dc output voltage that causes the vco to move frequency in the proper direction to achieve lock.

A phase-locked loop system has a basic property that must be taken into account when designing a frequency-synthesizer system. This is the amount of time required for the system to lock after a step in frequency. Generally, the lower the reference frequency, the longer the amount of time required to achieve lock. This is due primarily to the time constant of the components used to filter the dc output voltage. This explains the long time constant in some designs where the reference frequency is in the order of hundreds of hertz.

In one design the required output frequency was approximately 6 MHz for the transmitter and the desired channel

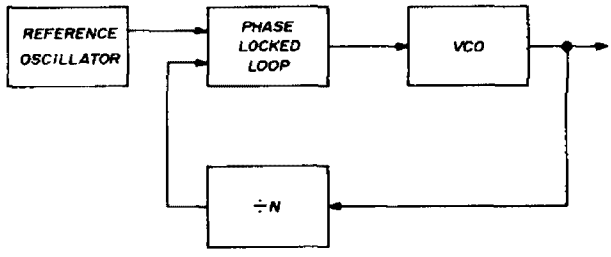


fig. 2. Phase-locked loop using a programmable frequency divider.

spacing, at 144 MHz, was 10 kHz. In this design, the 6-MHz signal was divided by the counter to produce an output of 416.666 Hz, which is equal to the reference. After multiplying the vco output by 24, these 416.666-Hz steps produced a 10-kHz change in frequency at 144 MHz. Because of the very low reference frequency, lockup time was approximately 3 seconds. Of course, this is much too long to be useful in a transceiver. This design also required special switching functions to achieve the required divide ratio.

The frequency synthesizer shown in fig. 3 is much the same as fig. 2, but an additional frequency divider has been added to the output of the vco and the reference frequency has been designated as 10 kHz. The variable counter is programmed so that each step in frequency produces a 10-kHz change in the vco output. This keeps the reference frequency high enough so it is easy to filter. It will also allow the use of easily available thumbwheel switches to provide direct readout.

The added divider is used to divide the output of the vco by an amount equal to the multiplication factor of the trans-

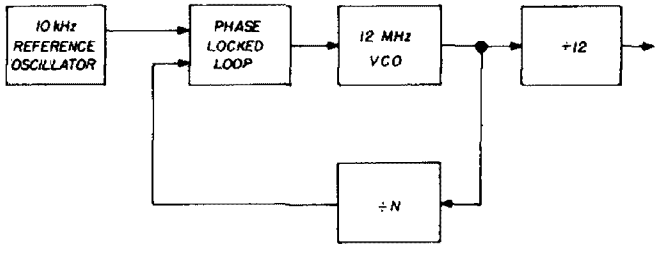


fig. 3. Phase-locked loop with an output frequency divider.

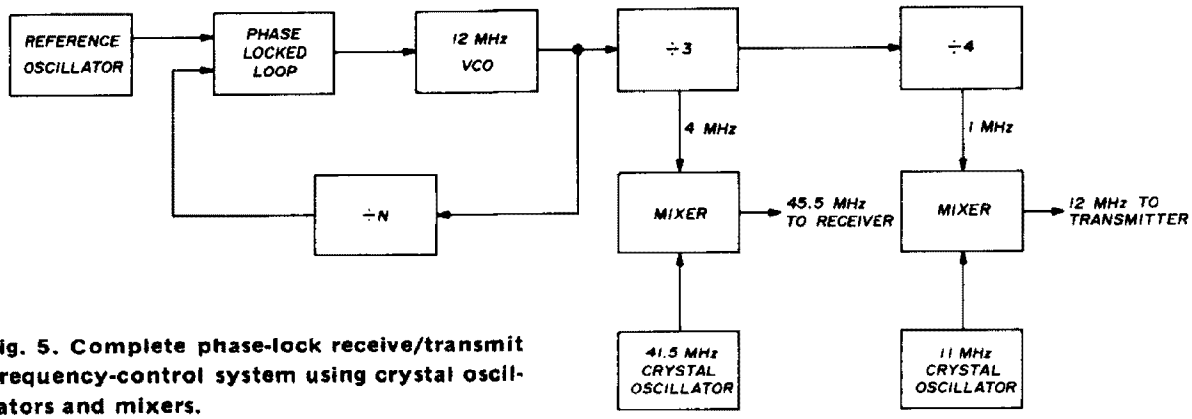


fig. 5. Complete phase-lock receive/transmit frequency-control system using crystal oscillators and mixers.

mitter. Assume the transmitter multiplies the crystal frequency by 12. Assuming a vco frequency of 12 MHz, the output is divided by 12, resulting in 1 MHz. This 1-MHz signal is mixed with 11 MHz, producing a frequency of 12 MHz. This is multiplied by 12 to 144 MHz.

If it is desired to transmit on 144.010 MHz, the programmable counter is stepped from the original 1200 to 1201. The vco frequency is now 12.010 MHz, divided by 12 to 1,000.833 Hz, mixed with 11 MHz to 12,000.833 Hz and multiplied by 12 in the transmitter to provide the final output at 144.010 MHz.

As can be seen from fig. 4, this system requires the use of a mixer and another crystal oscillator and doesn't provide for the receive function. However, a one-crystal system will be shown (the one crystal oscillator is that of the reference) that will provide both the transmit and receive frequency control and will be suitable for transceiver operation.

receiver control

The basic system shown in figs. 3 and 4 is applicable to any transmitter multiplication factor. Normally, the vco frequency should equal that of that original transmitter crystal which is being replaced by the PLL. This system was developed for a General Electric TPL rig. Although the TPL originally used 6-MHz crystals in the transmitter with a multiplication of 24, 12-MHz crystals seemed to work just as well and produced adequate deviation.

However, to include receiver frequency control in the system requires the addition of another frequency divider (see fig. 5). The required divide-by-12 (for the transmitter) is replaced by divide-by-3 and divide-by-4 circuits connected in series. This technique provides a control frequency for the receiver. Using a first receiver i-f of 8.7 MHz, the required injection frequency for reception on 144 MHz is 135.3 MHz. This, divided by three (the multiplication factor in the receiver) is 45.1 MHz.

In fig. 5 the basic 12-MHz vco frequency is divided by 3, providing an output at 4 MHz. This 4-MHz signal is mixed with a 41.1-MHz crystal oscillator to obtain the 45.1-MHz receiver injection frequency. The system of fig. 5 provides complete frequency control for both transmit and receive with one programmable divider and three crystals. How-

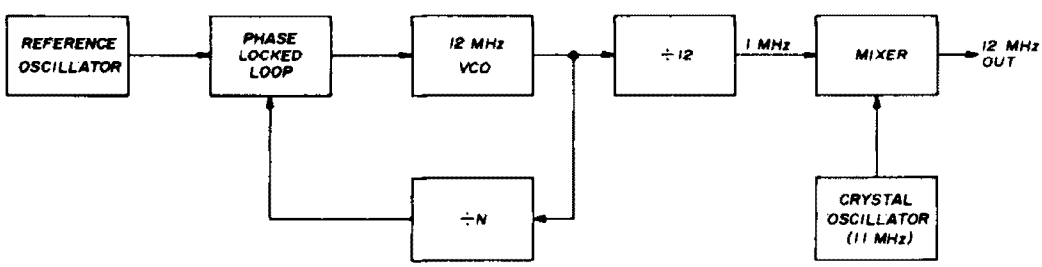


fig. 4. Phase-locked transmitter frequency-control system providing 10-kHz steps on two meters.

ever, it is possible to develop a one-crystal system that has the same stability as the reference oscillator. The system of fig. 5 is as stable as the two crystal oscillators because their instabilities are additive. However, because of some of the odd intermediate frequencies in some receivers, it may be necessary to use this technique to avoid a complicated frequency-division system.

It is important, before proceeding, to choose the frequency of the reference oscillator. A 100-kHz crystal will provide adequate stability, especially if it's mounted in a proportional oven. Remember that, when choosing a reference oscillator frequency, to a point the higher the frequency of the crystal, the better the stability. Although it depends on the particular crystal and its characteristics, normally, if a 1-MHz crystal is used, its stability will be better than that of a 100-kHz crystal.

The utmost in stability should be sought because any frequency error in the synthesizer will be multiplied by the transmitter and receiver multiplier chains. Also, it can be shown that drift in the reference oscillator will produce a frequency error in the output of the vco. The drift is a ratio between the divide program of the programmable counter and the reference error. By measuring the reference error, it is possible to predict the vco error. The often-used phrase, "good enough for ham use," is not good enough. Most amateur repeater and mobile installations are commercial quality, and we should strive to keep it that way.

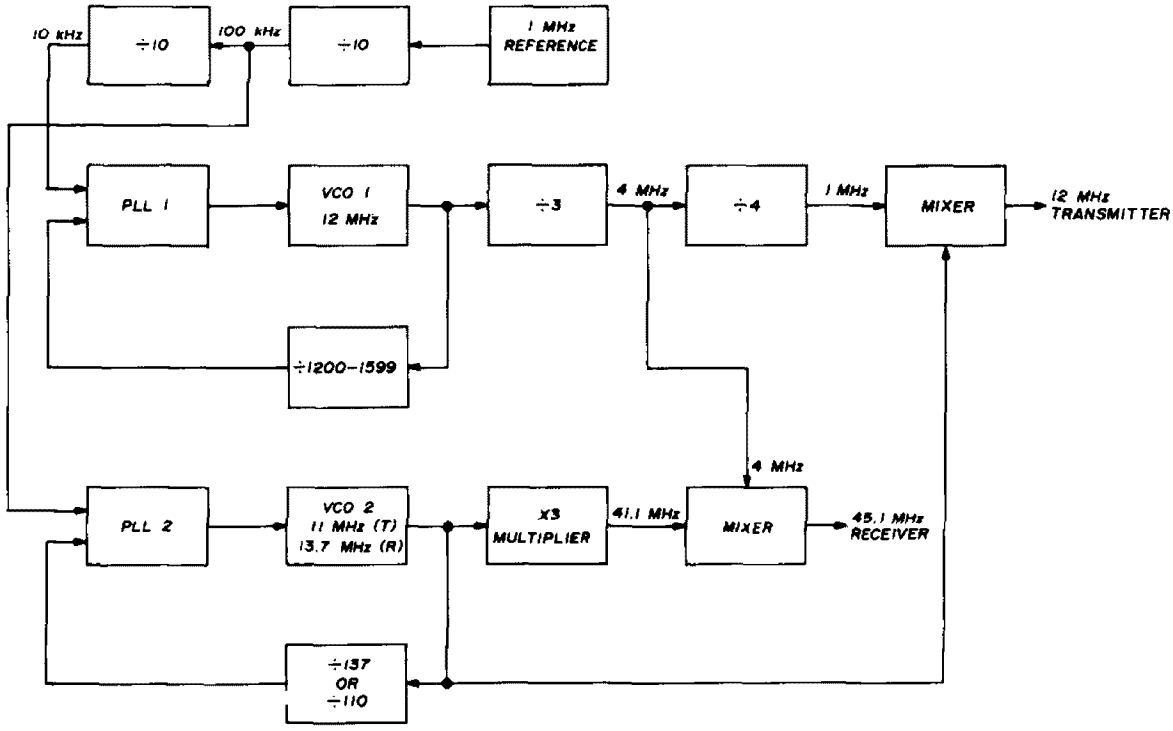


fig. 6. Transmit/receive frequency-control system using two phase-locked loops. Output frequency stability is equal to that of the reference oscillator.

one-crystal phase-lock system

The one-crystal phase-lock frequency-synthesis system in fig. 6 provides complete receiver and transmitter frequency control. A 1-MHz reference oscillator has been chosen for this circuit with two divide-by-10 ICs. This provides outputs at 10 and 100 kHz. The 10-kHz signal is applied to the main synthesizer loop (loop 1); the 100-kHz signal is applied to a secondary loop (loop 2). The secondary loop replaces the added crystal oscillators shown in fig. 5. This means that the stability of the system is the same as that of the 1-MHz reference oscillator.

To understand the operation of the

two loops, start with the transmitter. The frequency of vco 1 is 12 MHz followed by a frequency division of 12. Loop 2 is programmed to 11 MHz and applied to a mixer along with the output of loop 1. The output to the transmitter is 12 MHz.

In the receive mode vco 2 has a frequency of 13.7 MHz which is multiplied by three to 41.1 MHz. This signal is mixed with the 4-MHz output from loop 1 to provide a receiver injection frequency at 45.1 MHz.

As was mentioned previously, this system was developed for the GE TPL rig. To adapt this frequency synthesis system to other frequency ranges, you will have to do a bit of pencil pushing, changing the basic frequencies and dividing/mixing schemes. This system can, however, be used as is with a receiver having a 10.7-MHz first i-f, although it will only cover 146 to 148 MHz.

programmable counters

The programmable counters used in loop 1 of fig. 6 are programmed from 1200 to 1599, corresponding to a frequency range of 12.000 to 15.990 MHz. The 10- and 100-kHz counters in fig. 7 are programmed with BCD-coded thumbwheel switches, providing direct frequency readout. The MHz counters are programmed by a rotary switch labeled 144, 145, 146 and 147 MHz. These four frequencies correspond to vco frequencies of 12, 13, 14 and 15 MHz.

The programmable counters in loop 2 (fig. 8) use the same arrangement. In this case, however, programming is controlled by internal gating and is not brought out to the front panel. The counters in loop 2 are programmed for 137 on receive and 110 on transmit, corresponding to 13.7 and 11.0 MHz, respectively, for vco 2.

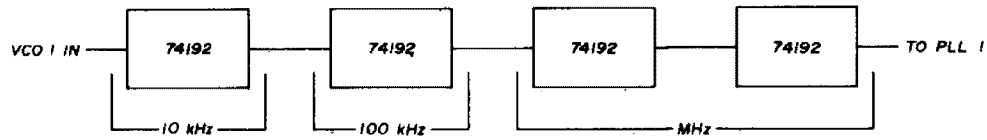


fig. 7. Programmable counters used in loop 1 of fig. 6 to divide by 1200 to 1599.

When using these counter ICs remember that the input programming terminals will rise toward the supply voltage if they are unused or left open. To preserve the noise immunity of the counter, which is especially important for mobile operation, these inputs should not be left floating or open; tie them to V_{cc} or ground, depending on the requirements of the system.

The control system shown in fig. 9 uses NAND gate switching and places all the counter inputs at the proper level. This control system also provides for transmit/receiver switching of the synthesizer oscillators for split-frequency operation.

The inputs to gates A1 through A4 and B1 through B4 in the control system are connected to V_{cc} through 4700-ohm resistors. The output of the BCD thumbwheel switches RX-S1, RX-S2, TX-S1 and TX-S2 are connected to the inputs of the A and B gates as shown in fig. 9. The wipers of the switches are connected to the C gates.

The push-to-talk circuit shown is for a grounded PTT system. That is, when you push the PTT switch to transmit, the PTT line is placed at ground. One input of gate

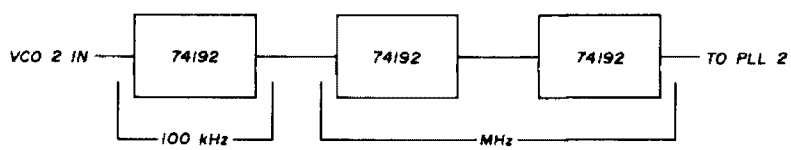


fig. 8. Programmable counters used in loop 2 of fig. 6 to provide divide by 137 (on receive) or divide by 110 (on transmit).

C1 is tied to the PTT line (V_{cc} on receive); the other C1 input is connected to V_{cc} through a 4700-ohm resistor. With both inputs high, the out is low, effectively grounding the wiper of the receiver thumbwheel switches.

One input to gate C2 is connected to thumbwheel switches are set to read 94. Therefore, 1, 2 and 8 of RX-S1 are open and 4 is grounded through gate C1. In RX-S2, 2 and 4 are open and 1 and 8 are grounded through C1. With 4 grounded, gate A3-A is grounded, making its output high. All other A gates have both inputs

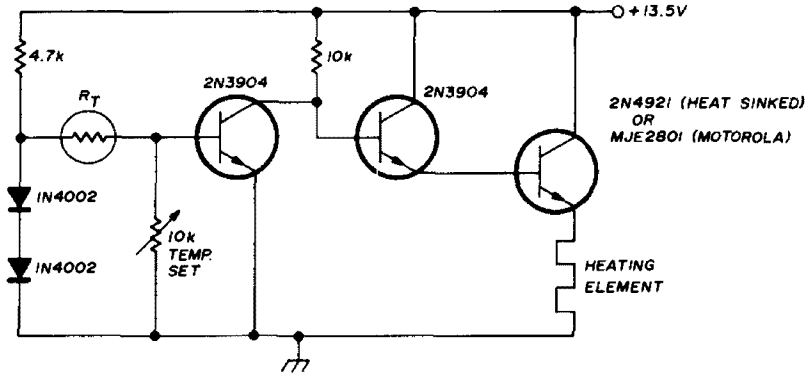


fig. 10. High-stability proportional oven control for the reference oscillator crystal. Thermistor R_T has 10k resistance at 25° C. Heating element is made from wire from an old crystal oven (see text).

the output of C1 and the other input is connected to V_{cc}. When the output of C1 is low, the output of C2 is high. Since the output of gate C2 is connected to the wipers of the *transmit* thumbwheel switches, their outputs are high during receive, regardless of their setting, and have no effect on gates A and B.

For example, assume you want to receive on 146.94 MHz. The receiver high, making their outputs low. Programmable counter PC1 is now programmed for 4. Since 1 and 8 of RX-S2 are grounded, counter PC2 is programmed for 9.

The remaining counters in loop 1 (PC3 and PC4) are programmed to 144, 145, 146 and 147 MHz with a rotary switch mounted on the front panel. Gates C3 and C4 control programmable counters

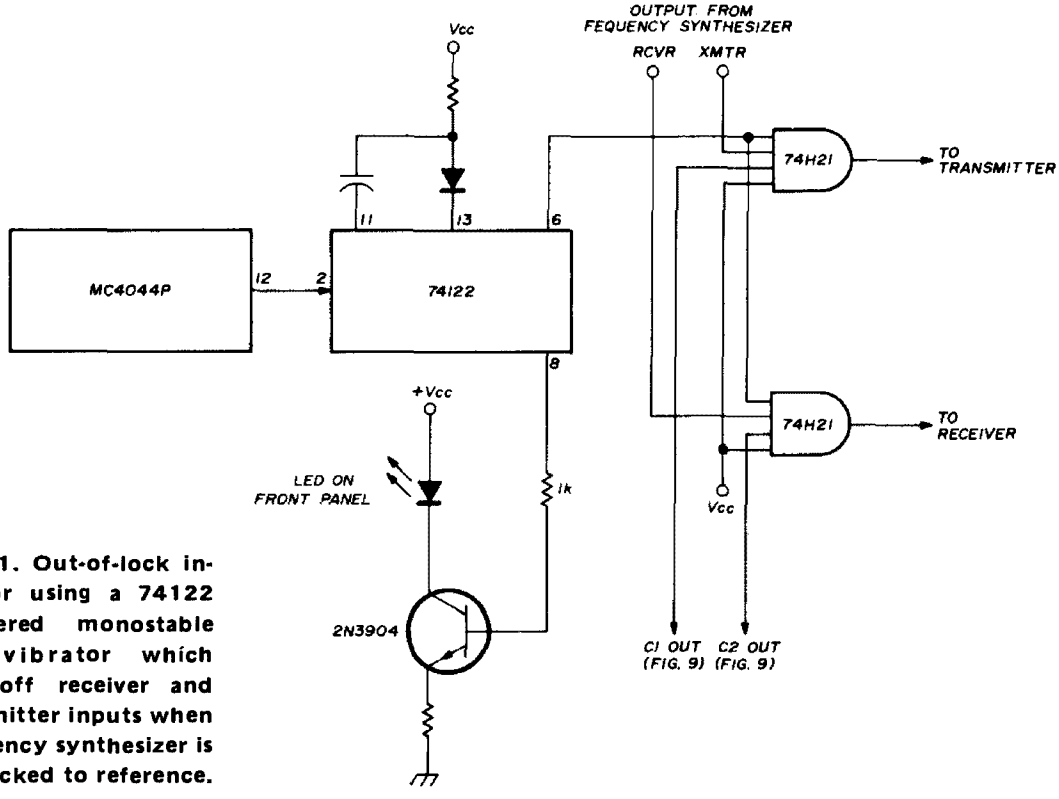


fig. 11. Out-of-lock indicator using a 74122 triggered monostable multivibrator which cuts off receiver and transmitter inputs when frequency synthesizer is not locked to reference.

PC5, PC6 and PC7 in loop 2 and program them for 137 on receive and 110 on transmit.

If the PTT applies a positive voltage to the system it is necessary to remove the 4700-ohm resistor at the input to C1 and install a 5-volt zener diode from this point to ground. It is also necessary to reverse the C1 and C2 wiper connections to the transmit and receive thumbwheel switches (receiver switches go to the output of gate C2 and the transmitter switches to the output of gate C1).

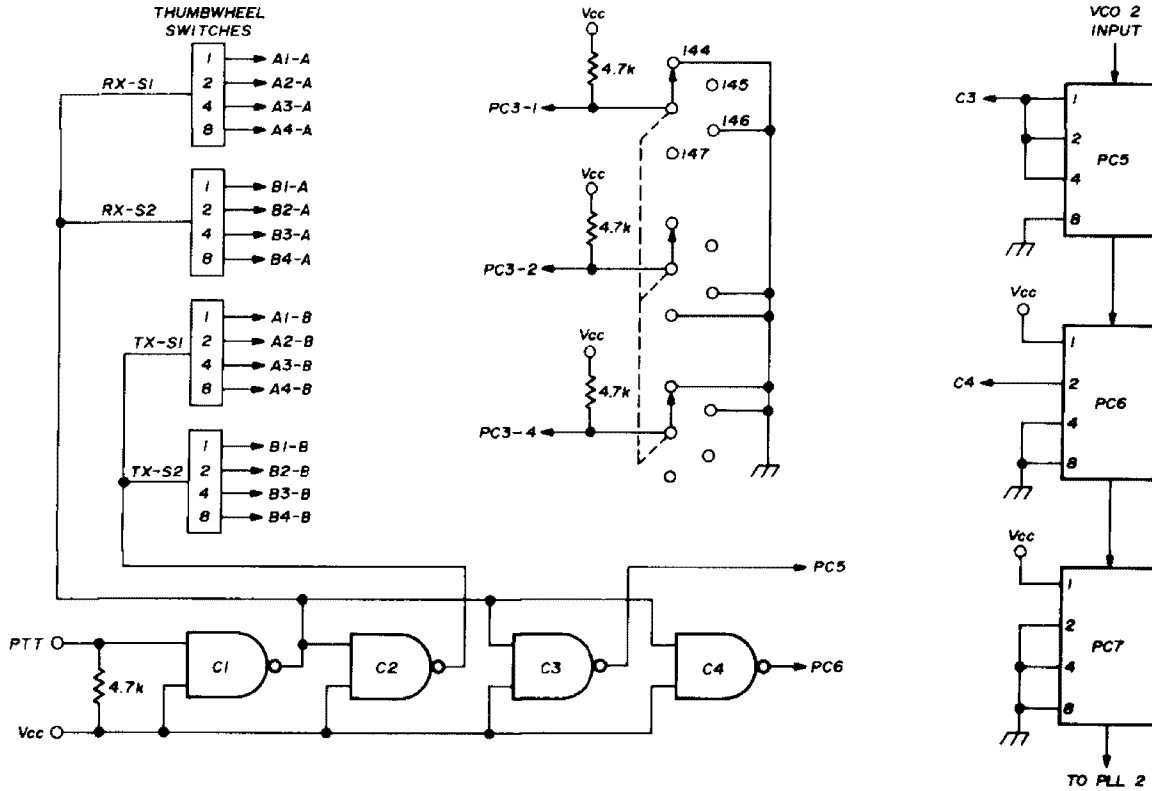


fig. 9. Control circuits for the programmable frequency dividers used in loops 1 and 2 (see fig. 6).

There are several programmable counter ICs that can be used in this system, but the least expensive of these is the 74192 up-down counter. This IC has the necessary frequency response and can be used in this circuit without additional gating.

proportional oven

For good frequency stability of the reference oscillator, a proportional oven is a must. Most thermostat-type items do not provide the close temperature regulation required for good frequency stability because they cycle with a 3- to 5-degree temperature differential.

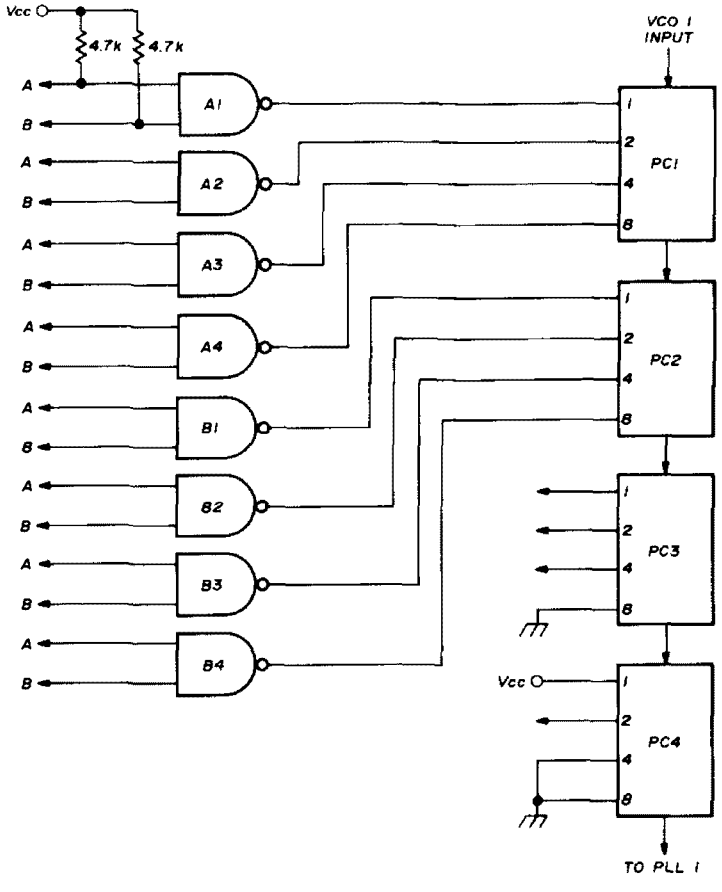
The simple proportional oven control circuit shown in fig. 10 has good temperature stability. The resistance element, R1, should be chosen to draw from 800 mA to 1 ampere when the oven is full on. I used the resistance wire from a 12-volt Motorola gold-type oven. The current through the heating element will be reduced to a much lower level when the oven has reached its operating temperature.

To reduce the sustaining current through the heating element a good layer of insulation should be installed around the oven. One way to do this is to use a small can to contain the crystal. The can can be wrapped with an insulating layer of fiberglass tape with the resistance wire wound on top of the tape.

The resistance wire should be non-inductively wound. To do this, fold the wire in half, and at the loop or fold-back end, attach a wire to the top of the can with a small piece of tape. Separate the two wire ends, and keeping them evenly spaced, wrap them around the can, working toward the bottom. After the resistance wire is in place, attach insulated

leads to the two ends and wrap the entire assembly with another layer of fiberglass tape.

This assembly is placed in a second, larger can. This second can should be big enough to contain from 1/8- to 1/4-inch of insulation such as sheet foam. The thicker the insulation, the better the oven and the lower the sustaining current.



The thermistor should be mounted inside the first can, as close as possible to the crystal. You may want to mount two crystal sockets in the can — one for the crystal and one for the thermistor. The thermistor can be installed on a base cannibalized from another crystal.

other circuits

The choice of a mixer circuit is left to the builder. I would prefer a dual-gate mosfet such as the Motorola MPF-121, followed by an amplifier with a drive control in the emitter circuit. The drive control is adjusted to the point of adequate drive across the desired frequency range; any overdriving of the circuit will result in undesired spurious output. The multiplier, mixer and amplifier should all have double-tuned tank circuits to help reduce spurious outputs and enhance broad-banding.

The phase-locked loop ICs I used for my breadboard experiments were Motorola MC4044Ps. These devices perform well, and with adequate filtering, the noise level was quite low as was any 10-kHz reference energy on the output. In addition, the MC4044P lends itself to building a simple out-of-lock indicator **(fig. 11)**.

Pin 12 of the MC4044P is always high when the loop is locked, but produces negative-going pulses when the loop is unlocked. If this output is applied to a triggered monostable multivibrator, such as the 74122 or 74123, and the time constant of the multivibrator is longer than time between pulses, the output can be used to control a gate that will cutoff the output of the synthesizer until the loop is locked.

I used a Motorola MC4024P IC for my vco. This is a dual voltage-controlled multivibrator with a typical maximum operating frequency of 30 MHz. Since the MC4024 is a dual unit, it can be used for both vco circuits required in the one-crystal frequency synthesizer. The maximum frequency I could obtain from the units was 24 MHz, but that may have been because the bread-board wiring capacitance limited the upper frequency. No interaction was observed between the two oscillators.

oscillator stability

Several times I have mentioned the importance of providing a very stable reference oscillator. One way to obtain excellent stability is to use WWVB as a reference. WWVB in Boulder, on 60 kHz, is stable to better than 1 part in 10^{12} per day and frequency accuracy is usually better than 2 parts in 10^{11}. Since vlf propagation is primarily ground wave, signal strength is adequate in most parts of the country for use as a reference.

The receiver for WWVB could be quite simple. The tuned rf amplifier could use IC amplifiers and tuned circuits made from toroid pot cores or old 50-kHz i-f cans retuned to 60 kHz. After amplification, the WWVB signal should be passed through a limiter circuit to remove amplitude variations and noise. The signal would then be squared up with a Schmitt trigger to be used as a reference with the phase-locked loops (see fig. 12).

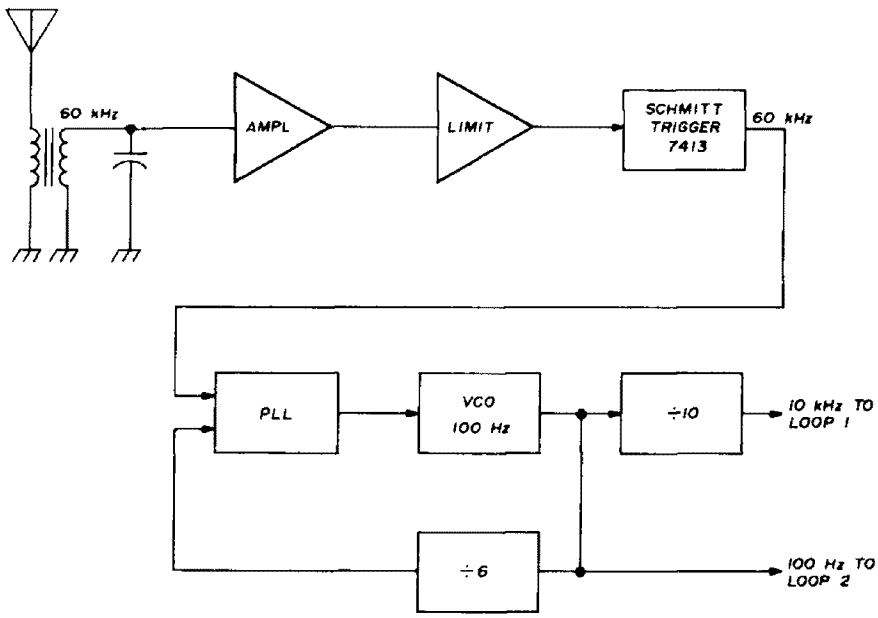


fig. 12. Using the received WWVB signal as the reference in a two-meter frequency synthesizer.

Because of the reference frequencies required for this synthesizer, a 60-kHz vco would not be usable. However, if the vco were at 100 kHz and divided down to 60 kHz for input to the PLL, the 100-kHz signal could be used for loop 2 and divided by 10 for loop 1. With the vco running at 100 kHz, it should be stable enough to withstand momentary losses of signal, such as might be encountered in mobile operation, without appreciably drifting off frequency.

Since WWVB goes off the air at specified times for maintenance, it would be necessary to provide a secondary oscillator that could be switched in, when necessary. The PLL is not perfect, nor can it maintain a perfect lock or phase relationship, so stability would not be as high as that of WWVB itself, but even so, it should surpass almost every commercial standard.

summary

I hope that this brief discussion of frequency synthesizers will stimulate interest in these very useful devices. I realize that this is not the ultimate synthesizer, but then, no one piece of equipment ever is. I would be happy to hear from anyone with similar ideas, or with suggestions for improving the basic design.

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ham radio



low-power vhf dummy load

The tobacco load — a low-vswr dummy load that will handle 30 watts of rf

Dave Sprague, WB9DNI, 78 Paddock Street, Aurora, Illinois 60538

Here is a one-evening project that will provide a very useful accessory for your shack — a 50-ohm dummy load capable of dissipating more than 30 watts of continuous power with a vswr below 1.1:1 through 220 MHz and typically, below 1.5:1 at 450 MHz.

The rf load is neatly packaged in a 7-ounce pipe tobacco can, filled almost full with transformer oil. The load itself consists of ten 510-ohm (5%) 1-watt carbon resistors parallel-mounted between pieces of copper weatherstripping. An alternate design approach is to use 2-watt resistors for a 70-watt load. However, there is a tradeoff with this approach as the vswr at 220 MHz may be as high as 1.8:1.

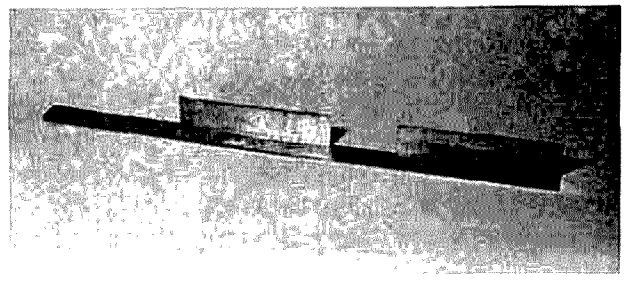


fig. 2. Copper strip used for mounting the 5% 1-watt resistors used in the rf load.

construction

Two types of 7-ounce tobacco cans are readily available from your pipe-smoking friends. They differ in the type of lid. The older type uses a paint can type lid; the second, and newer, type uses a screw-on lid. Although either tobacco can will work, I used the screw type for this project.

The metal which these cans are made from is soft enough that a scribe can be used to punch the necessary pilot holes when mounting the connector and vent tube in the lid. First, test the tobacco can for leaks by filling it with water. If any leaks are found, seal the seams of the can with a thin coat of RTV or similar oil-resistant sealer.

I tried several materials for the element mounting strips, but I found ordinary copper-clad weatherstripping to be electrically suitable, inexpensive and easy to form. Cut the strips to the specifications of fig. 1. Fold the larger strip as shown in figs. 2 and 3.

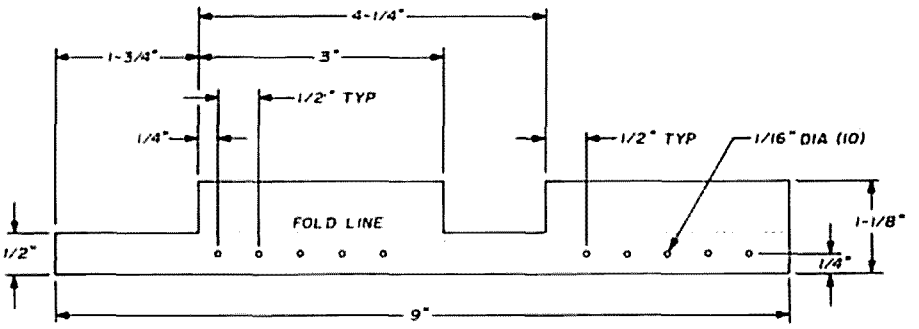
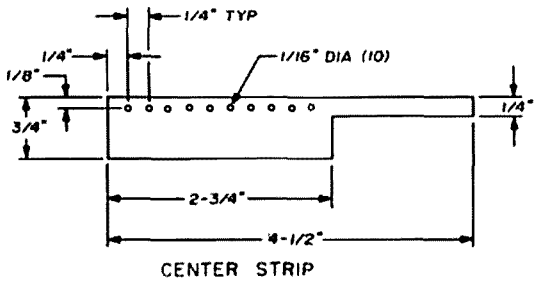


fig. 1. Layout of the copper strips used in the construction of the 30-watt rf load.



When the copper strips are completed, lay the smaller strip before you, as pictured in fig. 1B. Insert the first five resistors in alternate holes beginning with the hole closest to the right end. The end of the resistor body should be flat against the strip, but do not cut the soldered lead off at this time. A vise is a great help to keep the resistors properly aligned while soldering.

Next, turn the copper strip over and insert the remaining resistors in place with their bodies similarly aligned. Don't be stingy with the solder here — a good electrical connection is more important than looks. You should now have a unit which resembles fig. 4.

Cut the soldered lead of each resistor flush with the solder joint. Trim the free lead of each resistor to approximately ½ inch. Insert the center strip and resistors into the outer strip by carefully spreading the edges of the outer strip (see fig. 5) and guiding each resistor lead through the

Completed rf load. Small tube to the left of the coaxial connector allows heat expansion of the cooling oil.



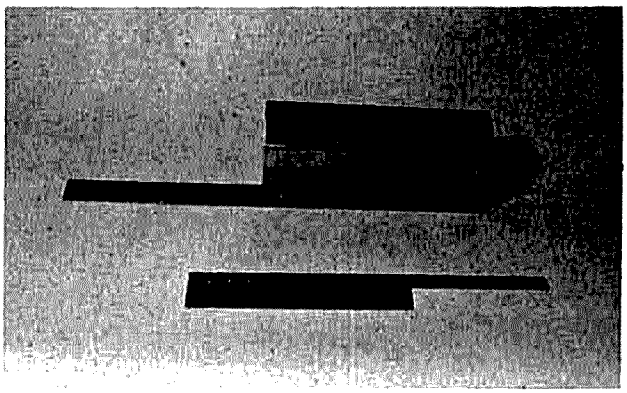


fig. 3. Copper strip is formed into a closed U-shaped box as shown here before installing the center strip and 1-watt resistors.

appropriate hole. Solder the free lead of each resistor to the outer copper strip, keeping the strip edge as close to the resistor body as possible. Trim the excess lead flush with the solder joint.

Solder the back and bottom edge of the outer copper strip as shown in fig. 6. When soldering, keep the center and outer strips separated. After verifying a resistance of 51 ohms between the strip tabs the load element is ready to mount in the tobacco can.

Mount a female SO-239 uhf coaxial connector through the lid of the can. I used pop-rivets to mount the connector but screws and nuts will work just as well if screw length is kept to a minimum.

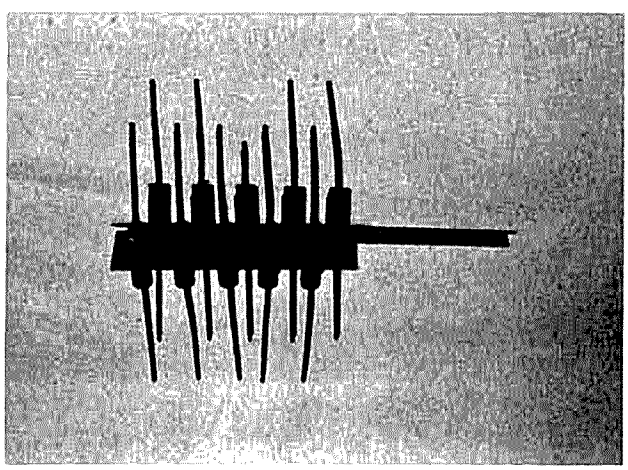


fig. 4. First, solder the resistors to the center strip. This center strip is then mounted in the U-shaped box as shown in fig. 5.

Solder a 3/4-inch piece of 1/8-inch OD copper tubing through the can lid so that the bottom of the tube extends to approximately 1/16-inch inside the lid. The load element should be mounted on the lid so that the bottom just touches the tobacco can bottom when the lid is secured (see fig. 7). There is sufficient tab length to fit almost any can type. Cut the excess from each tab, and solder the center tab to the SO-239 connector; solder the outer tab directly to the lid.

operation

Fill the tobacco can about two-thirds full of a transformer oil.* Insert the load

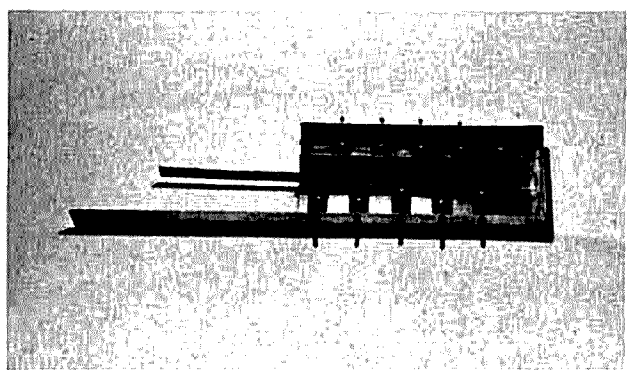


fig. 5. Install the center copper strip with resistors into the U-shaped box.

element and secure the lid. Remove the lid and note the highest point the oil reaches on the resistance element. Add enough oil to just cover the entire element. This leaves sufficient space for expansion as the oil heats. Secure the lid on the can tightly, and tape the seam.

Use a through-line type rf wattmeter when testing the tobacco load. Or, use a vhf reflectometer between the transmitter and the load. Keep the power low for the initial trial. There is nothing to tune or adjust. If the vswr is high, recheck the

*A small quantity of transformer oil can usually be obtained from your local, friendly, public utility company. Use caution if a substitute oil is used — the dielectric constant of vegetable oil, for example, is not consistent.

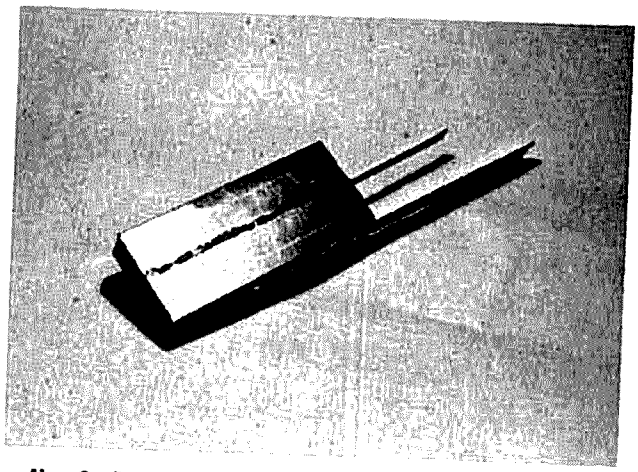


fig. 6. Solder the resistor leads and edges of the center copper strip to the U-shaped copper box.

dimensions of the load element and all solder joints.

I found that the tobacco load was able to withstand several rf overloads for short periods of time without damage to the load element. The thirty-watt rating is on the conservative side. With proper care this small load will pay for itself many times over.

ham radio

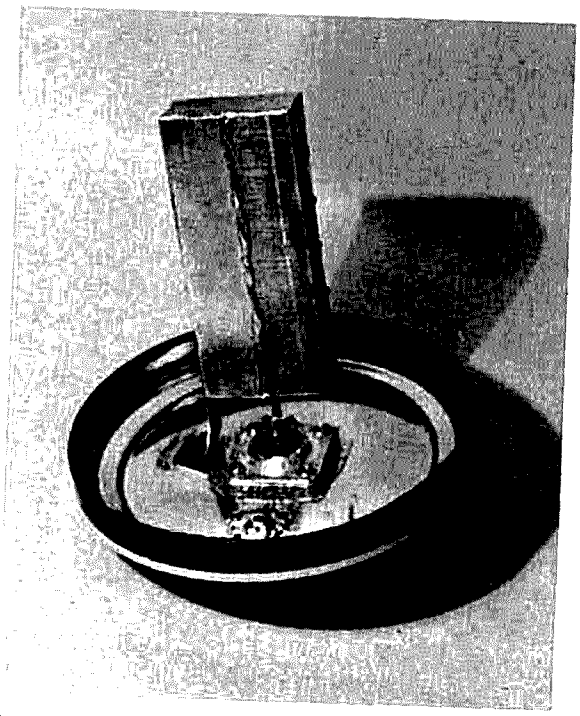


fig. 7. An SO-239 coaxial connector is soldered to the lid of the tobacco can and the load element is installed.

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vertical monopole log-periodic antennas for 40 and 80 meters

How to build low-cost log-periodic antennas for complete coverage of the amateur 40- and 80-meter bands

George E. Smith, W4AEO 1816 Brevard Place, Camden, South Carolina 29020

In previous articles I have described some homebrew, dipole-type log-periodic antennas, including antennas for 20, 15 and 10 meters[1,3] and 40, 20 and 15.[2] The log periodics discussed in these articles have provided gain on the order of 8 to 13 dB, depending on their length, their height above ground, the number of elements and the apex angle.

While testing some log periodics for use on the 7-MHz band I decided to try to lower the vertical angle of radiation to obtain improved, long-range communications performance. One way this could be accomplished would be to raise the antenna to 1/2-wavelength above the ground (or better still, to raise the antenna to 140 feet, a full wavelength). However, this would require at least two, and preferably four, 70- or 140-foot masts, so the cost would be very high. My 40-meter log periodic, at 50-feet above the ground, was as high as I could go, since I was using pine trees as supports.

Another method of lowering the vertical angle of radiation would be to use a vertical log periodic. Although I had previously tried a 10-, 15- and 20-meter log periodic in the vertical configuration by suspending it from some tall pine trees, this would be impractical for the 40-meter antenna because the longest element is 70-feet long. However, it

would be possible to do this with a single 75-foot mast.

A possible alternative would be to use quarter-wave vertical elements working against ground or a ground plane. This is similar to the vertical monopole log-periodic configurations used by some commercial stations for long-haul high-frequency circuits.

40-meter monopole log-periodic

I determined from a scale drawing that, by using only five elements, it would be possible to suspend a monopole 40-meter log periodic from a single catenary line strung between two high pine trees. With this construction, the horizontal radials would be about 20-feet off the ground (see fig. 1). A similar approach was used for a 75-meter log-periodic which I built later. The swr for both antennas was less than 1.5:1 over the entire design frequency range.

If this antenna could be installed a few feet above a salt marsh or a high-conductive ground, the radials could probably be eliminated. Some commercial monopole log-periodics use buried ground radials in place of the counterpose shown in fig. 1.

It should be noted that the ground radials become decreasingly shorter from the rear to the front of the antenna. They should be about the same length or slightly longer than their companion vertical elements.

Construction of the center insulators for this antenna is shown in fig. 2. These insulators are made from 1/4-inch thick lucite, 5/8-inch wide and 3-inches long. The two outside holes are for the radials; the holes on 1-1/2-inch centers are for the two-wire feeder. The center hole is for the quarter-wave vertical radiating element.

The vertical radiating elements and the ground-plane radials are fed by an open-wire feeder located at the bottom of vertical radiators. The transposed feed system required for log periodics is shown in fig. 1. A 4:1 balun is connected to the short-element end of the open wire feeder; a coaxial transmission line is used from the balun to the transmitter.

The driven or active element (element number 2) is approximately one quarter wave-length from the balun feed point. The quarter-wave line provides an impedance step-up. Since the impedance at the lower end of element number 2 is quite low, the quarter-wave transformer steps up the impedance to 200 to 300 ohms. The 4:1 balun provides a fairly good match to 50-ohm coax. The swr from 7.0 to 7.3 MHz is comparatively flat and less than 1.5:1 over the entire 40-meter band.

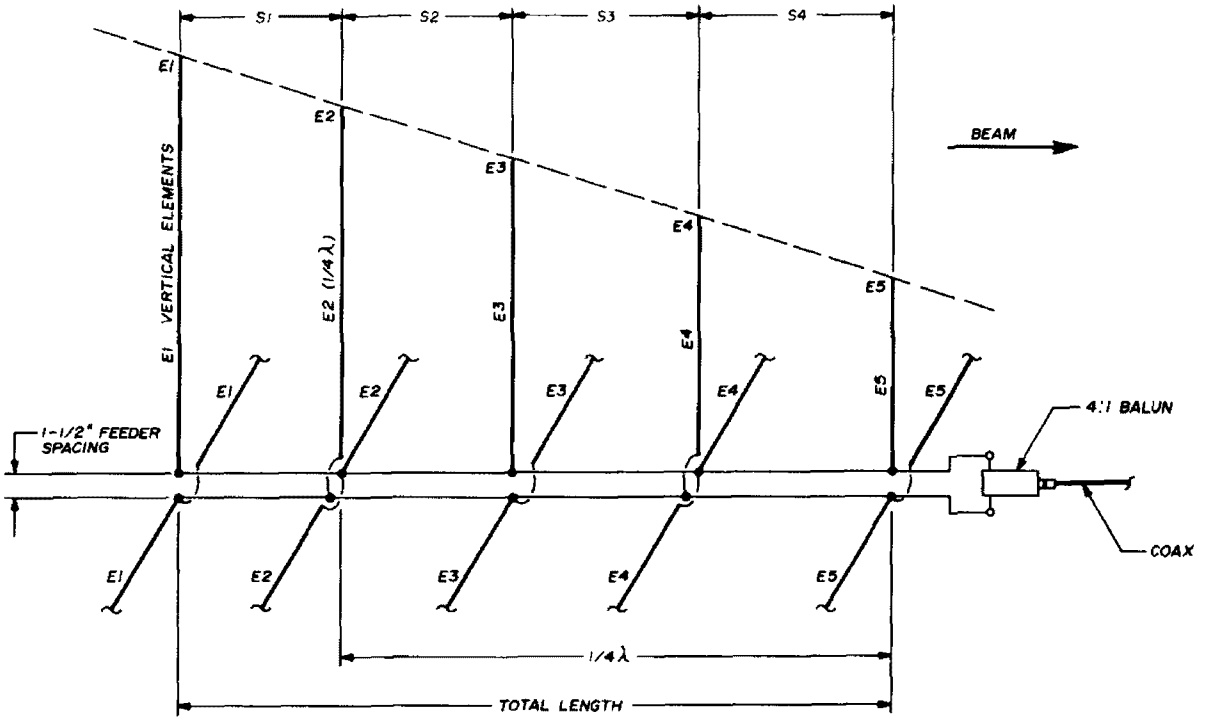
A 1/4-inch nylon line was used as the catenary to suspend the five vertical elements. This line was stretched between two supports as shown in fig. 1. The ground-plane radials were placed about 20-feet above the ground because clearance was necessary for some fruit trees which are located under the antenna. The outer ends of the radials are about 10-feet above the ground and are anchored to stakes or convenient posts or trees.

A total of 13 ground radials were used with this antenna: one to either side of the five vertical elements, plus two to the rear (element 1) and one to the front (element 5). Monofilament fish line (40-pound test) is used as guys and insulators for the radials.

performance

At distances greater than 400 miles, this antenna showed 8- to 10-dB gain over a commercial 40-20-15-10 trap vertical which works quite well for DX. The vertical is mounted on a roof 35-feet above ground and has all the radials, including four for 40 meters. The extra radial for 40 is necessary to lower the swr on that band.

Signals off the back of the log periodic



dimensions (feet)

	3.5-4.0	3.8-4.0	7.0-7.3	14.0-14.35
E1	70	65	35	17.5
E2	67	62	33	16.5
E3	58	55	28	14.0
E4	50	45	24.5	12.25
E5	43	40	20	10.0
S1	30	26	14	7.0
S2	27	24	13	6.5
S3	24	23	12	6.0
S4	19	18	9	4.5
total length	100	91	48	24
mast height	80	75	50	30
pole height	45	40	25	20

were just about the same as the trap vertical. As would be expected, stations just past the skip zone reported better signal strength from my half-wave doublet, 50-feet in the air (due to its higher angle of radiation). At distances greater than about 400 miles, however, the log periodic started paying off. This greatly enhances 40-meter operation during the daylight hours. It was difficult to do much testing after dark because of the heavy QRM from foreign broadcast stations.

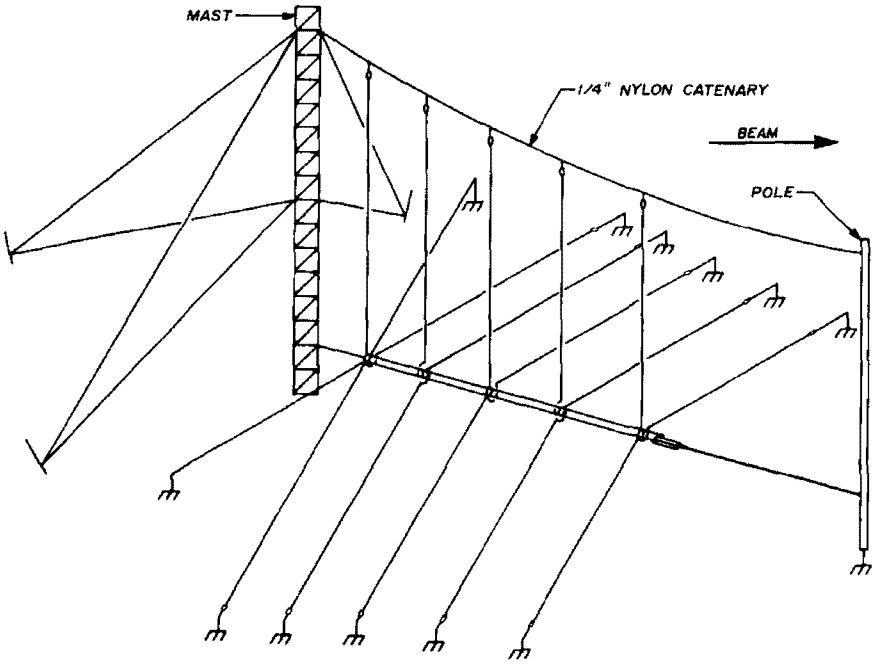


fig. 1. Five-element monopole log-periodic. Elements and radials are made from no. 15 aluminum wire. The 3.8-4.0 and 7.0-7.3 MHz designs (columns 2 and 3) have been built and tested. The swr was less than 1.5:1 over the entire frequency range. The designs for 3.5-4.0 and 14.0-14.35 (columns 1 and 4) have not been tested.

On reception, however, the vertically-polarized log periodic is much more susceptable to man-made noise than the horizontal doublet. As with any vertically-polarized, high-frequency antenna, it should be located as far as possible from industrial plants and busy streets. My antenna is about a block from the nearest street and is comparatively quiet except when cars come into my driveway.

80-meter log periodic

Since the 75-meter log periodic was erected only temporarily for test purposes, the ground radials and the open-wire feeder were only 3- to 4-feet above the ground. For a permanent installation this height should be increased to 10- to 15-feet. Since some of the radials have rf current circulating through them, the increased height above ground would prevent accidental contact with the radials. Also, this increased ground-to-radial height would increase the effective height of the antenna.

This antenna was originally designed to cover the entire 3.5- to 4.0-MHz band, but I discovered that there wasn't sufficient mast height for the rear element, which was 70-feet long. I decided to build a monopole log-periodic to cover only 3.8 to 4.0 MHz; this reduced the length of all elements by about 5 feet.

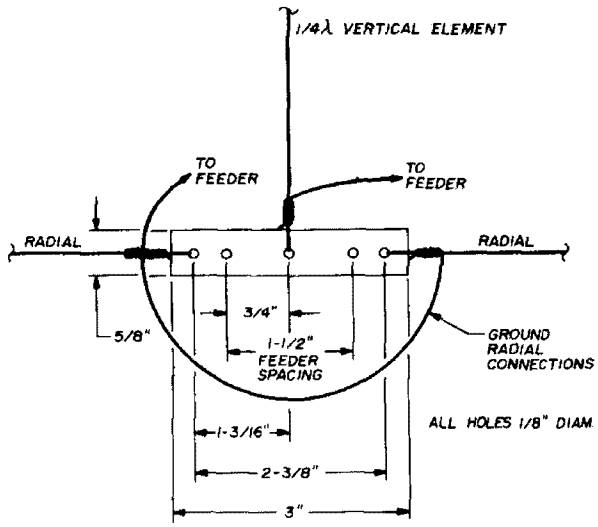


fig. 2. Construction of the center insulators. Material is 1/4" thick lucite. Connections to the open-wire feedline are shown in fig. 1.

The overall length from the rear to forward element is 91 feet. A minimum width of 135 feet is required at the rear and 80 feet at the front for installing the radial ground-plane system.

After the antenna was completed, I ran a series of vswr tests to determine if the swr was greater than 1.5:1 at any point in the band. The vswr readings were as follows:

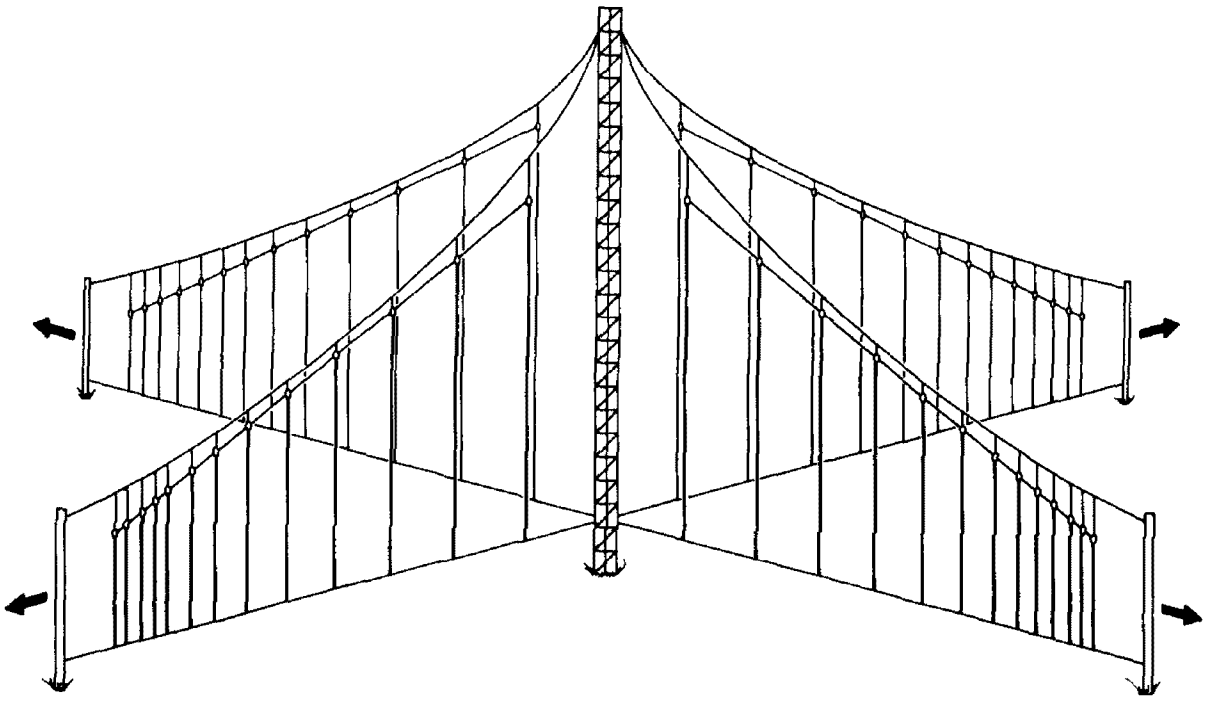


fig. 3. Suspending four vertical log-periodics from a single mast to obtain complete 360° coverage.

4.0 MHz	1.25:1	3.7 MHz	1.1:1
3.9 MHz	1.4:1	3.6 MHz	1.2:1
3.8 MHz	1.2:1	3.5 MHz	1.2:1

Although the swr is relatively flat over the entire 80-meter band, it is doubtful that the antenna has much directivity at the lower end of the band since it was designed to cover 3.8 to 4.0 MHz. On-the-air tests were confined to the upper 2 MHz of the band because the dipole I use for comparison purposes is limited to this meter bands, but further tests are needed to confirm this.

I was unable to obtain any data on front-to-back ratio as few stations were received off the back of the antenna. Side attenuation seems about the same as other log periodics I've used.

construction

The cost of assembling this antenna is quite reasonable, considering it will pro-

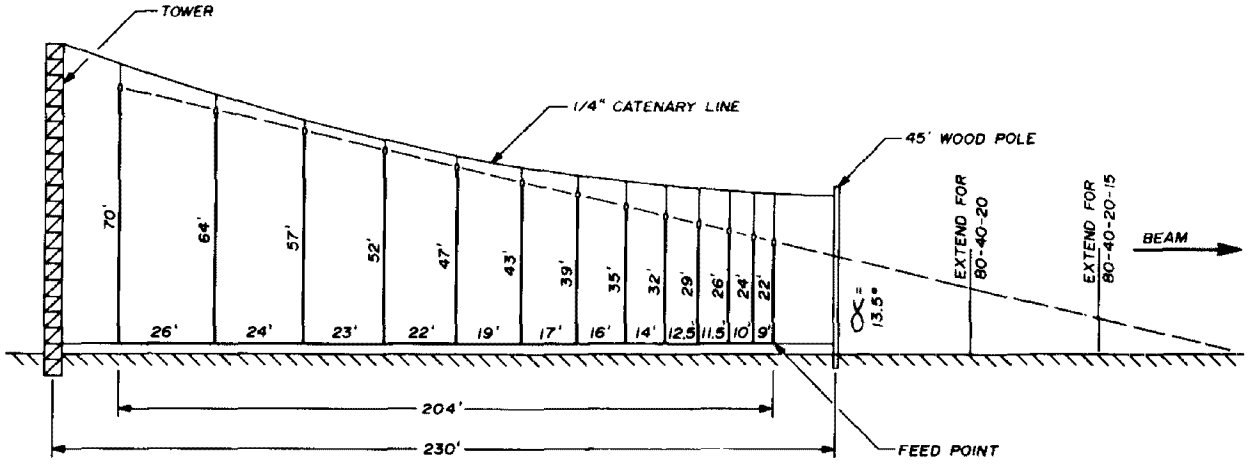


fig. 4. The vertical log-periodic can be extended to cover all frequencies from 3.5 to 7.3 MHz. This log-periodic uses 13 elements and requires a mast separation of 230 feet.

frequency range since the swr is greater than 1.5:1 below 3.8 MHz.

On-the-air tests indicated that the 80-meter monopole log periodic was best suited for distances greater than 500 to 600 miles during the daylight hours, and 1000 miles or more after dark. For shorter distances the horizontal dipole was usually one to three S-units better. However, at greater distances the lower radiation angle of the log periodic really began to do its stuff, and stations reported up to two S-units improvement with the log periodic.

For about one hour before and after sunrise and sunset, the log periodic is quite erratic, regardless of the distance. During these hours the dipole exhibited considerably less fading than the monopole log periodic. Also, the forward lobe of the log periodic appeared to be narrower than that experienced with log periodics on the 40-, 20-, 15- and 10-

vide up to 10-dB gain for DX. About 850-feet of wire is required for the elements plus another 185 feet for the open-wire feeder. I used number-15 aluminum wire which is manufactured for electric fences and can be purchased in quarter-mile coils for $8.70. Costs are also cut by using homemade lucite insulators and 40-pound-test monofilament fish line. The most expensive item in the installation is the 4:1 balun, which retails for $8.95. Total cost is about $25.00 (the 40-meter monopole log periodic is somewhat less, costing about $17.00).

other configurations

If space permits, two of these antennas could be mounted back to back, the rear elements suspended from a single mast. For more complete coverage three, or even four, beams could be supported by the same mast (see fig. 3). If you have enough real estate, the 80-meter log

periodic can be extended as shown in **fig. 4** to provide operation on both the 3.5- and 7.0-MHz bands. This two-band arrangement is about 205-feet long and would require a mast separation of 230 feet. A total of 13 elements are used in this antenna, about the minimum suggested for a log-periodic covering a 2:1 frequency range.

Since most amateurs aren't interested in the 3-MHz segment between the top 50-feet above the ground. The five quarter-wave monopole elements were suspended from the center feeder insulators. Since the rear element is 35-feet long, its lower end was about 15-feet above the ground (see **fig. 6**).

The object of this test was to elevate the current loop as high as possible, possibly lowering the vertical angle of radiation. Getting the current loop as high above ground as possible is similar to

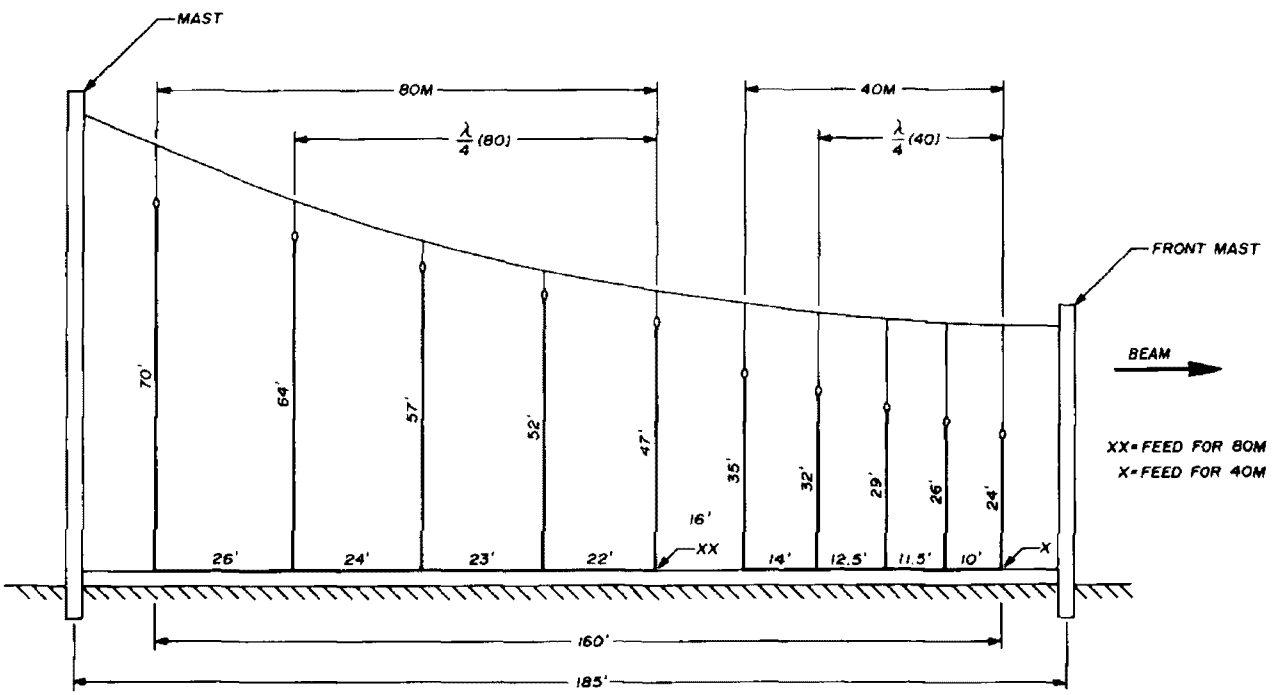


fig. 5. Breaking the log-periodic into two separate sections, one for 40 and one for 80, saves considerable space, but requires a separate feed system for each section. This antenna covers the frequency range, 3.5 to 4.0 MHz and 7.0 to 7.3 MHz.

end of 80 and the low end of 40, considerable space can be saved by breaking the antenna into two parts as shown in **fig. 5**. However, with this system two separate feedlines are required.

inverted monopole log periodic

After I completed the tests on the 40-meter log periodic shown in **fig. 1**, as an experiment I decided to try this antenna inverted, with the radials at the top. By using several trees as supports, I was able to suspend two nylon catenary lines to support the radials. This allowed the radials and the center open-wire feeder to be suspended approximately the old Bruce array which uses quarter-wave vertical elements in phase, spaced one-half wavelength.

Although time did not permit conclusive testing of this particular antenna, after a week of testing its gain seemed to be greater, or to increase, the more distant the station (over 500 or 600 miles). For stations just past the skip zone the horizontal dipole had the edge.

I had hoped to beam this experimental antenna to the west or northwest to run tests with W5s and W7s, since there seems to be a great deal of 40-meter activity there, but unfortunately, the trees I use for supports were all in the wrong places.

Therefore, I had to beam the experimental log periodic to the south. This limited the maximum distance to about 600 miles (Miami) as few stations further south were operating on 40 meters.

The forward lobe of this experimental antenna seemed to be narrower than that exhibited by the horizontal log periodic described in the previous article.[2] This is in line with other vertically polarized log-periodics I tried on the higher frequency bands which also had sharper forward lobes than their horizontal counterparts.

If the trees had allowed this "upside-down" monopole to be aimed to the northwest I'm sure I could have made some interesting tests. This configuration might also have possibilities as a low-angle beam for working DX on 80.

summary

The 40-meter log-periodic antennas are capable of 8 to 10-dB gain in the forward direction. Compared to the 3- to 5-dB gain provided by most 40-meter beams, the log periodic is worth consideration. The horizontal or inverted-vee log-periodic is suited for normal operation and compares favorably with a doublet except for its higher gain in the forward direction. The vertical monopole log-periodic is better suited for working DX.

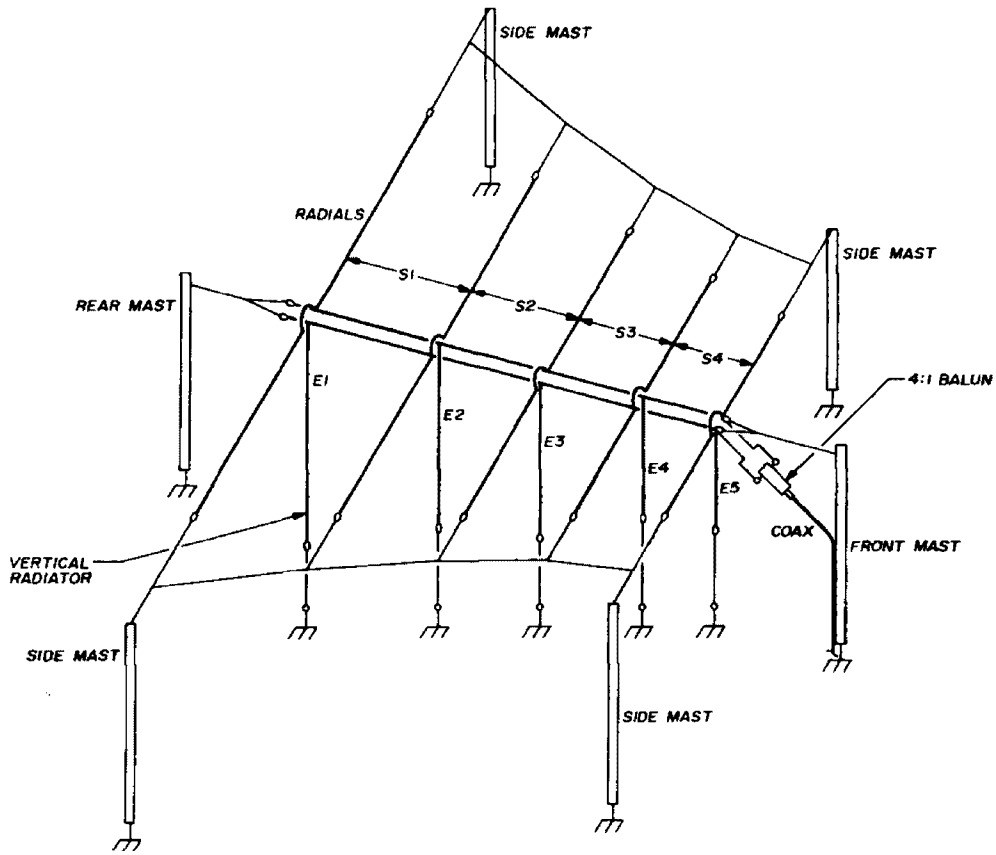


fig. 6. Inverted vertical monopole log periodic for 40 meters.

dimensions (feet) for 40 meters

E1	35	S1	14
E2	32	S2	13
E3	28	S3	12
E4	24.5	S4	9
E5	20		

references

1. G.E. Smith, W4AEO, "Three-Band High-Frequency Log-Periodic Antennas," *ham radio*, September, 1972, page 28.
2. G.E. Smith, W4AEO, "Log-Periodic Antennas for 40 Meters," *ham radio*, May, 1973, page 16.
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ham radio

noise reduction for CW reception

How to use threshold-gate/limiting and energy-integration to reduce noise interference on CW

John J. Duda, W2ELV, 6 Tuscarora Avenue, Geneseo, New York 14454

Much of the noise encountered in CW reception can be markedly reduced, and in many cases entirely eliminated, by appropriate processing of the received signal. It is well known that noise is a function of the bandwidth of the receiving system. Therefore, a narrow band-pass filter is one way to reduce noise. It is also well known that high amplitude impulse noise may be reduced by signal limiting.

In addition, there are other noise reduction methods, including using an audio filter followed by a detector which is used to operate a relay, which in turn keys an audio oscillator.[1] This technique eliminates all noise, but is not immune from keying by high intensity static bursts or by undesired signals on nearby frequencies. In addition to false keying there is the difficulty encountered by loss of "feel" for the signal. If the signal or receiver drifts off frequency there is no cue as to what direction you must tune to repeak the signal.

threshold gating

There are other methods of noise reduction, and, in general, best results may be obtained by using a combination of the various methods. The techniques discussed here are based on the fact that noise may be classified as either background hiss and hum, or as impulse noise. Usually, background hiss and hum is of lower amplitude than the CW note and may be eliminated by threshold-gating.[2,3] The basic threshold-gating circuit is reproduced in fig. **1**.

The post-gate filtering accomplished by the tuned circuit in series with the earphones not only reduces the distortion of the signal introduced by the gating process, it also reduces the amplitude of impulse noise which breaks through the gate. Thus, if the output signal at the earphones is fed through a second amplifier and subjected to a second threshold-gate, some of the impulse noise at the output of the first threshold-gate will not appear at the output of the second gate.

Noise is further reduced if the intermediate amplifier is frequency selective and peaked to the frequency of the post-gate filters. This process may be carried even further, through a third gate-filter, and beyond. The greater the number of stages, the greater the immunity to impulse noise.

The effectiveness of reiterative gate-filtering is a function of the gain settings of the intermediate amplifiers and the selectivity of the post-gate filters. Installation of a two-gate system is easily accomplished by substituting the headset in fig. **1** with a 3.2:2000-ohm matching transformer and using this to drive a power amplifier with another gate-filter in its output.

The 3.2-ohm winding of the matching transformer (an audio output transformer is satisfactory) replaces the headset while the 2000-ohm winding is terminated with a 2.2k resistor. This output is fed into the high impedance input of the audio amplifier. The amplifier must have sufficient power to operate the second threshold-gate filter; a minimum of one to two watts is acceptable. With this setup there will be a noticeable improvement over the performance of a single threshold-gate filter. The gain setting of the amplifier between the threshold-gates is fairly critical.

energy gating

Impulse noise may also be discriminated on the basis of energy content if it is possible to make reception contingent upon the energy content of the received signal or noise. Impulse noise tends to have less energy per unit time than a CW signal, and this difference can be put to use. Fig. 2 shows how an RC circuit can be used to integrate energy, providing a gate-control voltage to make signal passage dependent upon average energy content.

However, such a system appears to have two major weaknesses. First, it must truncate the received CW note, as passage of the note cannot occur until sufficient energy has accumulated. Secondly, it is vulnerable to post-signal impulse noise, as at the tail end of a note it is possible for impulse noise to integrate with the energy remaining from the CW note. Fig. 2 shows this.

The truncated CW note is not very serious, because if the output of the energy gate is fed to a tuned circuit, the

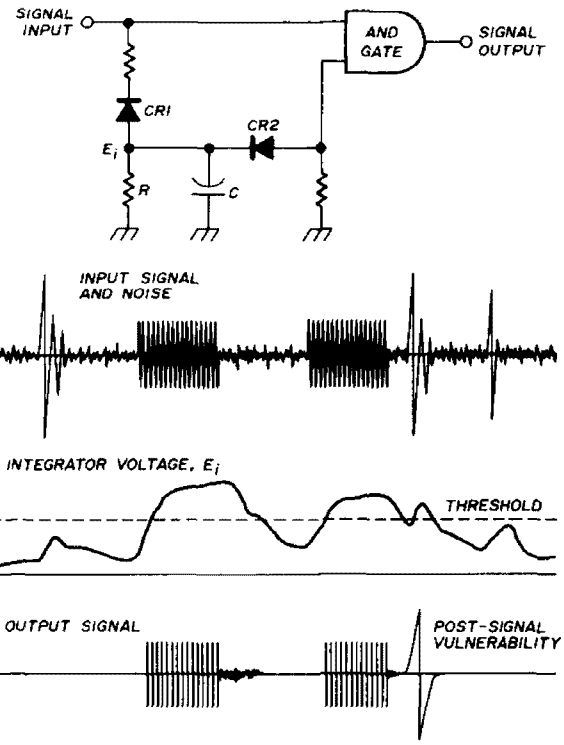


fig. 2. Energy-gating circuit. The RC integrator averages energy over time. The output of the integrator is used to operate an AND gate through a threshold device (CR2). Note the post-signal vulnerability.

Q of the circuit will tend to lengthen the CW character. With appropriate circuit parameters the two effects will cancel one another. However, there is no immediate solution to post-signal vulnerability. Fortunately, in practical applications it does not seem to be a very serious problem.

As can be seen, the value of the RC integrating circuit is fairly critical – the higher the value of the integrating capacitor, the greater the immunity to impulse noise and the greater the truncation of CW notes, as well as a greater post-signal vulnerability. The lower the value of the integrating capacitor, the opposite is true. Nevertheless, there is a range of RC values which give satisfactory results.

Experiments with energy gating

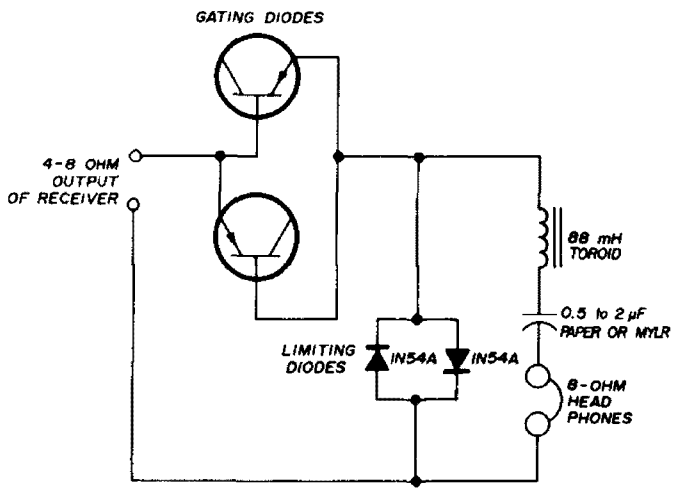


fig. 1. Basic threshold-gate/limiter circuit with post-gate filtering. The emitter-base junction of 2N414 transistors may be used for the gating diodes.

evolved from further developments in the threshold-gating circuit. In an attempt to increase the amount of post-gate filtering I built the circuit in fig. **3**. The two tuned circuits are brought into mutual resonance by placing an ac voltmeter (or dc milliameter with a diode and appropriate value of series resistance) across each capacitor and tuning the receiver through a carrier. Peak readings at each capacitor should occur at the same dial setting (same audio frequency). If there is a difference, mutual resonance can be established by adding padding capacitance across the capacitor which gives the peak reading at the higher audio frequency.

The circuits have sufficient Q to introduce a slight amount of ringing so a small amount of damping was introduced to reduce this. In the circuit in fig. **3** there is a diode and variable resistor in series across the second tuned circuit. The variable resistor is set to give the most pleasing note. Once this setting is determined you can measure the resistance of the variable and replace it with a fixed resistor of the same value.

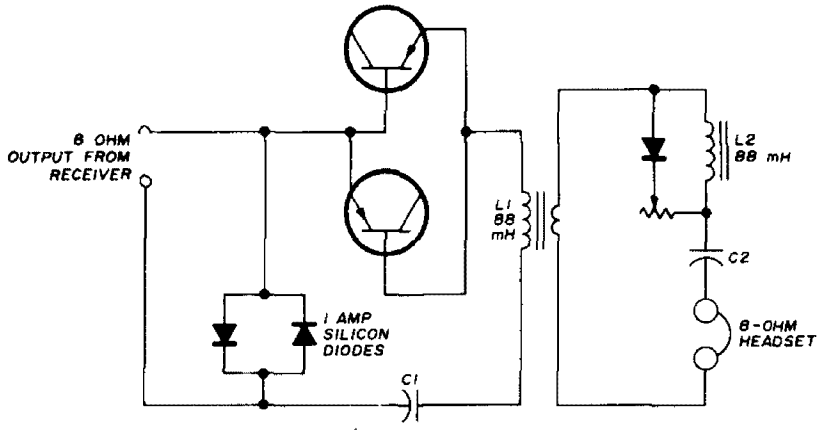


fig. 3. Threshold-gate/ limiter with two-stage, post-gate filter. Tuned circuits L1-C1 and L2-C2 are resonated to the same frequency. Link on L1 is 32 turns no. 28. Value for C1 and C2 falls in the range 0.5 to 2.0 μF; exact value is a function of the desired peak frequency.

combining the systems

An improved circuit that incorporates threshold-gating, post-gate filtering, energy gating and tone characteristic control is shown in fig. **4**. The Q of the first tuned circuit serves to multiply the driving signal to a level adequate for operating the integrating circuit (CR1, R1 and C1). Transistor Q1 serves a switching function; diode CR2 and the emitter-base characteristics of Q1 provide a threshold effect.

A variation of this circuit is given in **fig. 5**. Here, the resistor in series with the base connection of Q1 is eliminated and the keying voltage from the integrating circuit is made variable. This adjustment effects the keying characteristics of Q1 and there seems to be a critical setting at

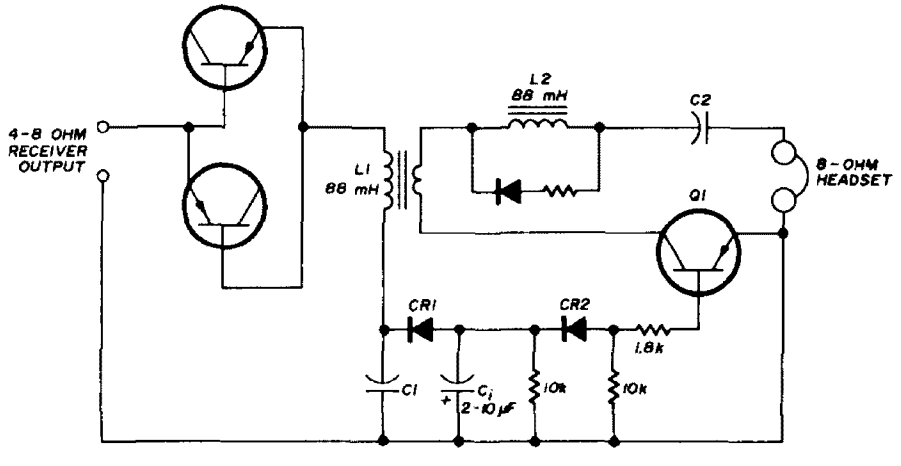


fig. 4. Threshold-gate circuit with both amplitude and average energy thresholds. C1 is the energy integrating capacitor; 2 to 10 μF provides satisfactory results. Diode CR2 provides threshold keying of Q1, which acts as a switch in the headset circuit. Link on L1 is 32 turns no. 28.

which the CW notes are more crisp, and hence, more readable than usual.

The circuits in figs. 4 and 5 provide complete elimination of some forms and amplitudes of impulse noise. Total elimination of impulse noise is not possible, however, if the circuits are overdriven or if the impulse noise is very severe. There is some similarity between energy-gating and that method which uses the received signal to key an independent audio oscillator. In energy gating it is the signal itself which is keyed. Thus, it is possible to discriminate between impulse noise and signal, or from one signal to another, when these happen to key the output. In addition, some feel of the receiver tuning is maintained.

To evaluate reiterative-gating and energy-gating I put together a two-stage system. The first threshold-gate filter was followed by a ten-watt hi-fi amplifier which fed the circuit of fig. 5. The receiver I used had a Q multiplier and an internal audio filter, and I found it necessary to lower the Q of the receiver's audio filter to avoid ringing (all tuned audio circuits were set to the same frequency).

With this receiving system I found that signals as close as 250 Hz could be completely separated. Radioteletype signals were received as two distinct signals, with a zone of silence in between. Most interesting, however, was the immunity to sideband energy from radiotelephone signals. Apparently, there was insufficient average energy to operate the energy gate. It was possible to tune through the 40-meter band at night and not be bothered by the international shortwave stations which operate there. Each of these signals were reduced to a carrier which occupied about 200-Hz of dial space. Beyond that, there was no indication of their presence.

summary

There is much that can be done to achieve crystal clear CW reception. The CW signal is amenable to various types of signal processing, as all you have to preserve is its timing characteristics. The circuits presented here are perhaps the most simple ways of obtaining threshold- and energy-gating. With more sophisticated circuitry it should be possible to achieve superior results.

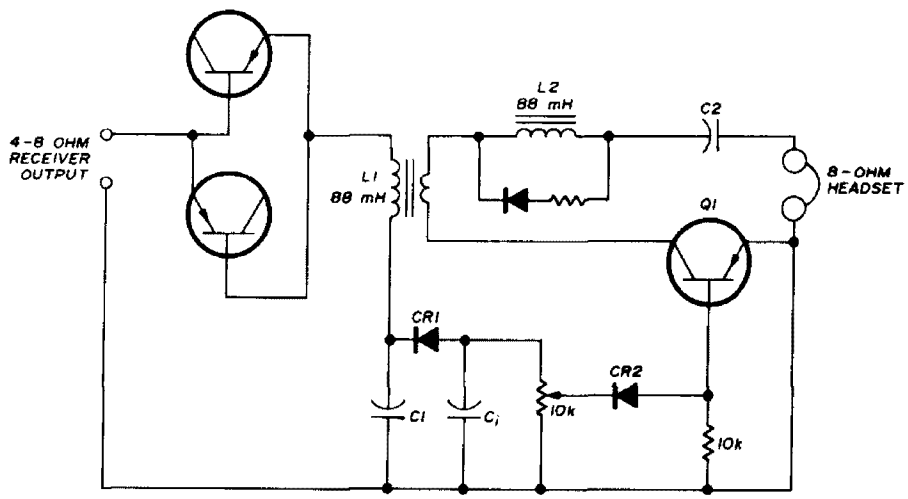


fig. 5. Alternate amplitude/energy gate circuit. Setting of the 10k variable resistor is critical for besting sounding CW note. Transistor Q1 and CR1 and CR2 are small-signal types. Link on L1 if 32 turns no. 28.

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2. J. Duda, W2ELV, "Threshold-Gate/Limiter for CW Reception," *ham radio,* January, 1972, page 46.
3. J. Duda, W2ELV, "Threshold-Gate/Limiter Addition," *ham radio,* May, 1972, page 59.

ham radio

two-capacitor transmission-line matching system

Eliminate antenna matching stubs with this simple two-capacitor matching system

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It is often desired to match an open-wire transmission line to a resonant antenna whose resistance is different than the characteristic impedance of the line. There are many sources which adequately describe how this may be done using shorted or open stubs, but very little is available on doing this with a combination of lumped constants and a transmission-line section. This article describes a convenient matching system using fixed capacitors connected across the open-wire line a specific distance apart, and shows how to find that distance.

The use of capacitors, particularly the relatively inexpensive fixed vacuum types still available on the surplus market, is attractive in this application because it eliminates the need for line sections hanging down, and facilitates experimental adjustment. Vacuum capacitors are weatherproof, probably more so than feeder spreaders. The losses involved are certainly comparable, if not less than with lines.

matching system

The system is essentially that of the double-stub tuner, except that for flexibility in achieving a match to a wider range of load (antenna) resistances, and to enable the use of capacitance values specified in advance, it will be assumed that the transmission-line matching section has a Z_0 different than the main transmission line to be matched.

The circuit analyzed is shown in fig. 1. The matching section has electrical length, θ, and characteristic impedance, Z_0. The two capacitors are represented as reactances X1 and X2, respectively, with X2 nearest the load (antenna) and X1 nearest the transmitter. Resistance R_a is the antenna resistance and R is the characteristic impedance of the main transmission line to be matched. Antenna reactance can be compensated for by simply moving the two capacitors along the line so that there is a section of line between X2 and the antenna.

The analysis proceeds along conventional lines assuming image-impedance operation and treating the line section and two capacitors as a network. The image impedances are unequal and made the same as the resistances R and R_a. The short-circuit and open-circuit impedances at each end of the line are computed, from which you can obtain both the ratio R/R_a and the product RR_a. Manipulation of these relationships then permits the

reactances to be found. The result, for capacitive X1 and X2, is

$$\frac{Z_o}{X1} = \frac{1}{T} \pm \sqrt{\frac{R}{R_a}} \sqrt{\frac{1}{T^2} + 1 - \frac{Z_o^2}{RR_a}} \quad (1)$$

$$\frac{Z_o}{X2} = \frac{1}{T} \pm \sqrt{\frac{1}{\frac{R}{R_a}}} \sqrt{\frac{1}{T^2} + 1 - \frac{Z_o^2}{RR_a}} \quad (2)$$

where $T = \tan \theta$. In these equations the same sign of the radical must be used in both equations. A necessary condition is that the quantity under the second radical be positive. For this to be true,

$$\sin \theta < \frac{\sqrt{RR_a}}{Z_o} \quad (3)$$

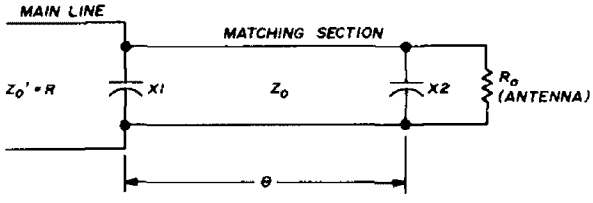


fig. 1. Basic two-capacitor line-matching system.

This places a limitation upon the length of line that can be used, or conversely, upon the range of impedances that can be matched. This is the same sort of limitation that applies in the case of conventional double-stub tuners where transmission line sections rather than capacitors are used. In this case, however, it was assumed that Z_0 can be established as something different than that of the main transmission line. If this were not done, $Z_0 = R$, and $\sin \theta < \sqrt{Ra/R}$ (from eq. 3). In practical situations, such as matching a 73-ohm antenna to a 600-ohm transmission line, you would find that the line section might have to be rather short, about 20° in this example. Therefore, it is necessary to retain the option of changing the Z_0, and, as will be seen later, you *must* have this option if you are to use capacitors of predetermined values.

With two capacitors whose values are fixed in advance, Z_0 becomes a variable and it is necessary to solve **eqs. 1** and **2** simultaneously to find Z_0 and $T = \tan \theta$. Fortunately, this is not too difficult and the result is

$$Z_o = \frac{R_o}{\sqrt{\left(\frac{m-\rho}{a}\right)^2 + 1 - \rho^2}} \quad (4)$$

$$T = \frac{R_o}{\rho Z_o} = \tan \theta \quad (5)$$

where

$$R_o = \sqrt{RRa}$$

$$m = \frac{R_o}{X_1}$$

$$n = \frac{R_o}{X_2}$$

$$a = \sqrt{\frac{R}{R_a}}$$

$$\rho = \frac{m - a^2 n}{1 - a^2}$$

Both m and n are positive for capacitive reactance. If the quantity under the radical turns out to be negative, you must choose other capacitor values and try again.

example

Following is a step-by-step procedure for finding the required Z_0 and length of the line section. There are some general rules that should be followed, however, to avoid having to repeat the calculations because one of the values turns out to be imaginary. First, select the capacitance values so that their reactance is somewhere around the characteristic impedance of the line to be matched. Second, install the largest capacitor on the side having the lowest impedance.

In this example it is desired to match a 73-ohm antenna to a 460-ohm line. Three

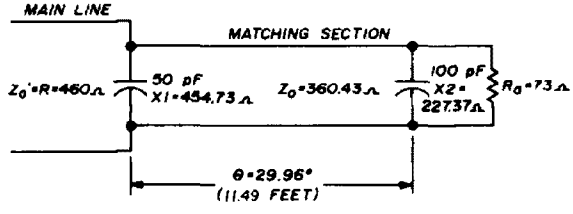


fig. 2. Two-capacitor system for matching a 73-ohm antenna to a 460-ohm line at 7.0 MHz.

50-pF vacuum capacitors are available. Since 50 pF has a reactance of 454.73 ohms at 7000 kHz, this would be a good value for the 40-meter band. In as much as three capacitors are available, two of them are placed in parallel and used for X2, nearest the low-impedance, or antenna side of the line. (If matching a higher impedance antenna, R_a greater than R, put the parallel capacitors on the line side, nearest the transmitter.)

To find the characteristic impedance of the matching section, Z_0, and the length of the matching section, it is necessary to go through the following calculations:

$$R_o = \sqrt{RR_a} = \sqrt{460 \times 73}$$

$$= 183.25 \text{ ohms}$$

$$a = \sqrt{R/R_a} = \sqrt{460/73} = 2.51$$

$$m = R_o/X_1 = 183.25/454.73 = .403$$

$$n = R_o/X_2 = 183.25/227.365 = .806$$

$$\rho = \frac{m - a^2 n}{1 - a^2}$$

$$= \frac{.403 - (2.51)^2 \times .806}{1 - 2.51^2} = .882$$

Now, using eq. 4, find the characteristic impedance of the matching section, Z_0

$$\frac{183.25}{\sqrt{\left(\frac{.403 - .882}{2.51}\right)^2 + 1 - (.882)^2}}$$

$$= 360.43 \text{ ohms}$$

Compute tan θ and find θ in the trigonometric tables

$$\tan \theta = R_o/\rho Z_o$$

$$= 183.25/(.882 \times 360.43) = .576$$

$$\theta = 29.96^\circ$$

When θ has been determined, the length of the line section can be calculated with the following formula

$$L = \frac{2.73}{f_{MHz}} \theta = (2.73/7)29.96 = 11.49 \text{ feet}$$

The completed matching system consists of a 360.43-ohm open-wire line, 11.49-feet long, with three vacuum capacitors, as shown in fig. 2. The line section can be made up applying the usual formulas for characteristic impedance,

$$Z_o = 276 \log_{10} \frac{2S}{d}$$

where S is the center-to-center spacing and d the wire diameter, in the same units. A 360.43-ohm line made from number-12 wire requires a spacing of 0.82 inches. If this is considered too narrow, then use a four-wire line for which

$$Z_o = 138 \log_{10} \sqrt{2} \frac{S}{d}$$

if the wires are arranged on a square, or

$$Z_o = 138 \log_{10} \frac{2D_2}{d\sqrt{1 + (D_2/D_1)^2}}$$

if the wires are arranged on a rectangle of sides D_1 and D_2. Diagonally opposing wires are connected together in a four-wire line of these types.

To compensate for antenna reactance, the distance of X2 from the antenna can be varied, or else a variable capacitor can be connected in parallel with X2 (or with one of the two capacitors removed) and tuned for minimum vswr on the main line. The best approach, of course, is to have the antenna cut to length at the desired frequency so that it is resonant and nonreactive.

summary

The same principles can be applied to other cases, say where you want to use inductance rather than capacitance. The same relationships apply, but with the sign changed on X1 and X2 (and hence, on m and n). A matching system is, after all, equivalent to a pi-matching network where impedance transformation takes place due to unequal image impedances of the network. You might even use the same general analytical approach to design a *three-capacitor* matching system that would then allow a predetermined Z_0 to be used for the line section.

ham radio

vari-Q filter

Remember the shunt Q multiplier? Here's how to use it in series with a narrowband mechanical filter

Stirling M. Olberg, W1SNN, 19 Loretta Road, Waltham, Massachusetts 02154

An affliction of the QRP bug resulted in a receiver accessory that is compact and has many features found in high-power stations. Receiver selectivity is always important; in QRP operation it often means the difference between completing a contact or no contact at all.

My receiver for QRP work is a superhet that has a 1.2-kHz mechanical filter and a Q multiplier. This combination reduces the overpowering effect of nearby signals and improves the signal-to-noise ratio. Used with a narrowband mechanical filter, the Q multiplier can be used to peak or null a signal within the mechanical filter passband.

selectivity

The combination of a 1.2-kHz mechanical filter and a Q multiplier offers advantages not found in many commercially built amateur receivers. The narrow bandwidth provided by the mechanical filter, in combination with the Q multiplier, provides discrimination between strong and weak signals; the Q multiplier allows you to peak or attenuate a signal across the mechanical filter passband.

Parallel-resonant circuits that have a high Q also present a high resistance, and thus a high impedance, which is subject to a reduction to a lower value due to the circuit to which it's connected. It's impossible to prevent this effect, but it can be reduced. In a practical parallel-resonant circuit, the impedance can be expressed as:

$$Z_r = QX \qquad (1)$$

where

Z_r = impedance at resonance (ohms)

Q = quality factor

X = reactance of C or L (ohms)

The result of **eq. 1** is that Q is degraded due to changes of reflected impedances caused by the Miller effect. Such degradations adversely affect the operation of the Q multiplier due to its method of coupling to the receiver. It

to understand the problem, is the 3-dB bandwidth of our parallel-resonant circuit. The bandwidth at the 3-dB points is:

$$BW = f_o/Q \qquad (2)$$

where

f_o = resonant frequency (kHz)
Q = quality factor

For a Q of 200 at 455 kHz ±3 dB, f_o = 455/200 = 2.2 kHz. The loaded Q is:

$$Q = R/X \qquad (3)$$

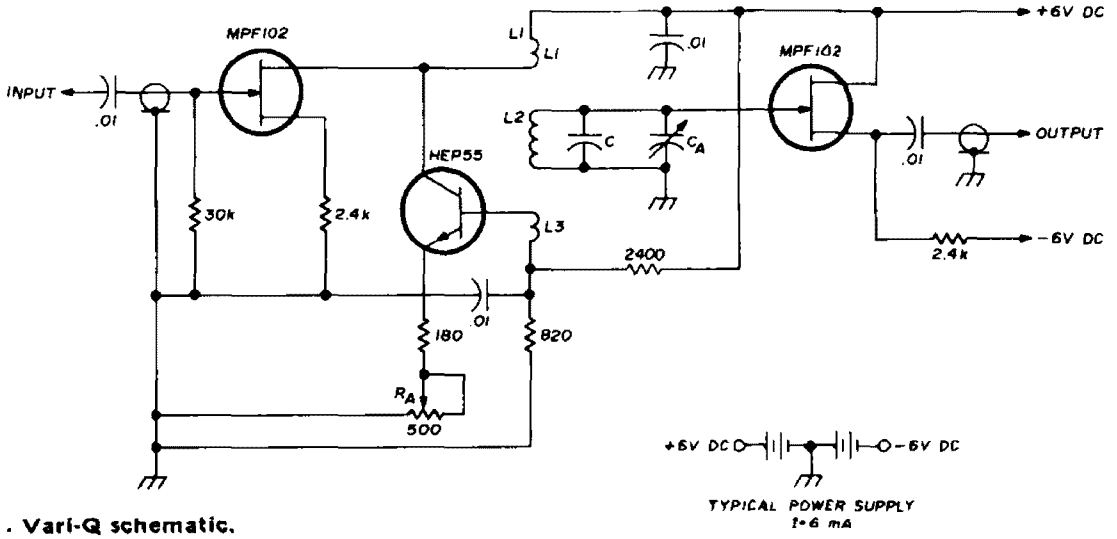


fig. 1. Vari-Q schematic.

became clear that isolation of input and output circuits would offer some improvement. This isolation was achieved in the circuit of fig. **1.**

The heart of the Q multiplier is an inductor with a high Q (typically of the order of 100). I found that the receiver circuitry associated with the Q multiplier degraded its Q so that full capability of the Q multiplier was not fully realized. A review of some basic theory will help to explain.

example. Let our inductor have an unloaded Q of 200. The inductive and capacitive reactance of the circuit are each 500 ohms. Thus Z_r = QX = 200 x 500 = 100k ohms, which is the impedance of the parallel-resonant circuit.

Another consideration, which will help

where

Q = quality factor
R = parallel load resistance (ohms)
X = reactance (ohms)

From **eqs. 2** and **3** it can be seen that bandwidth is an inverse function of load impedance.

the Q multiplier

Now that we've reviewed the facts regarding inductors in parallel resonance, let's take a closer look at the Q multiplier. The Q multiplier is an amplifier that functions as a very high-Q circuit. It is connected in parallel with one of the receiver i-f stages. A parallel-resonant circuit has a very high impedance. Any signal passing through the i-f at the Q multiplier resonant frequency sees a very

high shunt impedance and is not attenuated, whereas signals near the Q multiplier resonant frequency appear on the skirt of the Q multiplier response curve and are shunted to ground. As the regeneration control is advanced to a point just below oscillation, the effective Q of the circuit is increased. Thus the peaking action can be increased to any desired level.

It's possible to attenuate adjacent ssb signals in the same manner, but a much lower regeneration level is required. In CW operation, another interesting effect is the inability of the Q multiplier to respond to static or key clicks; this is because the Q will not build up fast enough to respond to such impulses.

the vari-Q circuit

The circuit acts like a variable filter with a bandwidth adjustment. For this reason "Vari Q Filter" is the name given to the circuit. Sufficient gain is available to compensate for any losses in its addition to most receivers.

A review of fig. 1 shows that the input circuit impedance approximates the reactance of an i-f transformer. The input is coupled to the gate of a fet, which is directly coupled to the collector of an HEP55 or similar transistor. The HEP55 serves as the Q-multiplier regenerative amplifier.

The base of the HEP55 is connected to one of the secondary windings of the transformer (L3). Feedback is produced by coupling the output into the input of this amplifier via the tuned secondary, which has the high Q required. The input and output circuits are arranged to be completely isolated. Also, sufficient gain is available to allow *regeneration to* occur.

The usual Q multiplier circuit is coupled very loosely to the i-f input through a so-called gimmick capacitor (twisted lead). The gimmick must be connected between the output of the mixer and the first i-f input as shown in fig. 2A. The Vari-Q Filter differs in its installation — it is connected in series as shown in fig. 2B. When the Vari-Q Filter is added to the i-f strip, the loading effects are so small that it acts as a Q multiplier. Of more importance, the Vari-Q Filter retains these properties across the full passband of the mechanical filter.

It is rarely required to operate the Vari-Q circuit near the unstable level since extremely narrow passbands can't be used; so a lower level of regeneration is used, which offers greater stability.

It is possible to completely eliminate one of two CW signals by adjusting the tuning capacitor, CA, to recenter the passband of the Vari-Q Filter. A very small loading effect is reflected into the LC combination. The output of the source follower is coupled through a small capacitor to the i-f transformer, which serves as an input. The 2.4k source follower load resistor can be changed to match any required load simply by substituting the desired value. It is important to use a balanced power supply. Current drain is 6 mA, which can be supplied from a battery.

Table 1 is included to allow the home brew artist to scale off his own frequency requirements. The toroids and pot cores indicated can be substituted with any inductor having a Q of the order of 100.

table 1. Coil and capacitor data. All coils are wound with no. 32 enamelled wire.

frequency (kHz)	L1 L2 L3 (no. turns)	core type	C CA (pF)
50	16 130 40	Ferroxcube 1811CA250-387	3600 10-75
455	12 115 24	Amidon T44-15	470 7-45
1800	6 100 18	Amidon T44-15	75 3-30

construction

The Vari-Q filter is built on a small piece of etched circuit board employing "circuit stiks." The variable tuning capacitor can be found in surplus bins. Its small size allows it to be mounted to the board directly, which reduces lead length and allows the unit to be mounted on the receiver front panel. Next to the capacitor a potentiometer, used as the regenera-

tion control, was added to the front panel, giving full control to the filter circuit.

operation

For best results the regeneration control should be at a low setting, and the tuning capacitor should be at mid range. Tune in the desired signal. If interference is present, adjust the regeneration control until a slight "plop" is heard, but do not advance it further. Then tune in either the desired signal or null the interferring signal by adjusting the variable capacitor. If greater selectivity is desired, advance

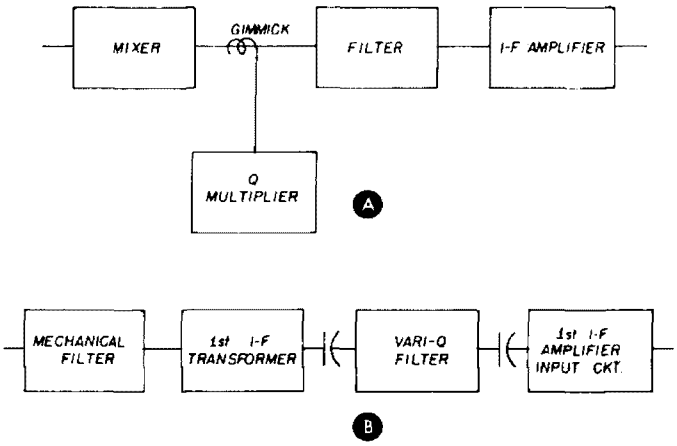


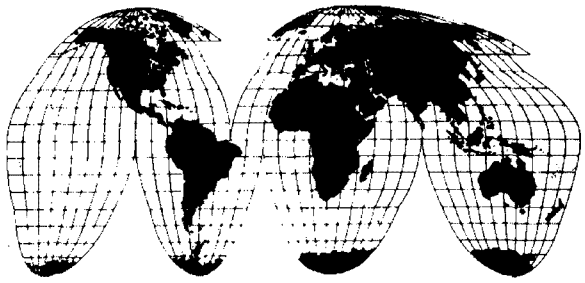
fig. 2. The usual Q multiplier is connected in parallel with the i-f as shown in (A). The Vari-Q Filter is connected in series as shown in (B).

the regeneration control until a ringing effect is heard. The amount of ringing that can be tolerated is a variable that depends on the operator. With experience and skill, it's possible to pull weak signals out of noise and interference even when background ringing occurs. The regeneration control has a fairly wide range and won't cause the unit to break into oscillation until the control is close to its maximum position.

Although the Vari Q was designed for a QRP receiver, it has also been incorporated into my 75A4 and UR388 receivers. Several others have also used it successfully in other receiver designs. The Vari Q will be manufactured either as a kit or as a complete unit. Details are available — send me a self-addressed stamped envelope.

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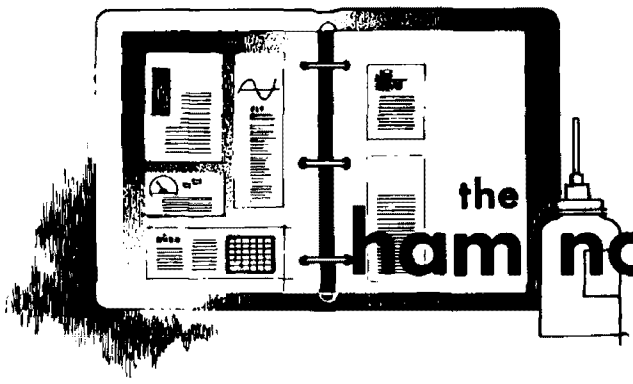
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the ham notebook

adding a cw filter

Many transceivers don't have sufficient selectivity for CW so a good audio filter must be added. Several good active filters are available, powered by a small battery. The filter should not be in the circuit while you scan the band or listen to ssb. Therefore, two switches may be needed; one for the filter and one for the battery. If wired as shown in **fig. 1**, a single dpst switch is all you need. In its *normal* position the filter is shorted out of the circuit and the battery circuit is open.

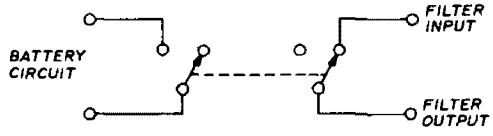


fig. 1. When installing a CW audio filter in your transceiver a single dpst switch may be used to turn off the power supply and by-pass the filter.

The filter may be added between the audio output of the receiver and an earpiece. If convenient, it is better to install it between audio stages, where the audio signal level is still low. In any case, make sure that shorting the filter input and output does not upset some dc circuit within the receiver or in the filter itself.

Because of the extreme selectivity of a good filter, an incoming beat frequency must be held to narrow limits. For example, a CWF-2 filter* has a center frequency of 750 Hz and a bandpass of only 80 Hz. A signal cannot peak as desired unless its beat note is near 750 Hz.

If your transceiver has *variable* offset control, the receiver can be tuned (for desired beat) without changing the transmitter tuning. My Swan 350C does not have such a control. Although adding the filter greatly improved CW reception, I decided to modify the set for maximum efficiency.

In the *receive* position a fixed 10-pF capacitor (C-140 on the schematic) is added across the carrier oscillator crystal. This creates a beat of 800 Hz. I replaced C-140 with a 25-pF variable. Now I can peak all signals, and can also chase a drifting signal without varying transmitter tuning.

Front-panel space is limited, so I removed the *rf gain* pot to make space for the 25-pF capacitor. Gain had always been left at maximum, anyway, so I didn't need the gain control. A fixed 10,000-ohm resistor replaced the pot.

The CWF-2 can be switched to several bandwidths, 180, 110 or 80 Hz. I prefer the latter at all times (except when scanning the band or listening to ssb).

I. Queen, W2OUX

*The CWF-2 CW audio filter is available from MFJ Enterprises, Box 494, State College, Mississippi 39762.

magnetic fields and the 7360

Having had occasion to work on several sideband transceivers using the 7360 magnetic-deflection mixer tube, I found these tubes much influenced by stray, external magnetic fields. The deflection plates inside the tube can also be permanently charged.

On one occasion I found the 7360 would not balance out the carrier in any position of the balance pot, while another

tube worked as it should. I tried a homebrew degaussing coil about two inches in diameter, dropping the tube through it with 60 cycle ac on the coil, thinking the field would become progressively weaker as the tube passed through. However, results were unsatisfactory and the tube still refused to balance.

With the tube in its socket, I passed a permanent magnet along the glass envelope from bottom to top. Still no improvement. When the magnet was reversed, however, the tube came to life and a good balance was obtained about one-third of the way through the pot rotation.

Watch where you lay the 7360 when out of its socket. A screwdriver on the bench may be magnetized, and alongside the 7360, it will do its worst. Even when testing the tube, don't put it down on top of the tester, the transformer may be mounted right underneath and not magnetically shielded. Also, watch out for steel tube shields for the 7360, they may become magnetized.

Eugene A. Hubbell, W7DI

finding square roots

The small, hand-held digital electronic calculator is showing up in large quantities in the marketplace, and every large discount department store has several from which to choose. These calculators are great to have in your home for every-day calculations or even electronic circuit design. Although most of the consumer-oriented calculators are excellent for the standard mathematical functions of addition, subtraction, multiplication and division, the square root function is missing. However, by application of the Mechanic's Rule,* extracting square roots may be obtained with greater than slide-rule accuracy. Although the answer has a small amount of error, the result may be used for most applications.

*James and James, *Mathematics Dictionary*, D.Van Nostrand Company, New York, third edition, page 233.

The Mechanic's Rule is as follows:

$$\sqrt{N} \cong \frac{\left(\frac{N}{\text{est}\sqrt{N}}\right) + \text{est}\sqrt{N}}{2}$$

where N is the number you want the square root of, and est ON is the estimated square root.

For example, find the square root of 54. Since 54 falls between 49 and 64 (square roots 7 and 8, respectively), the square root of 54 will be between 7 and 8, and probably, by inspection, around 7.3. Plugging these numbers into the Mechanic's Rule:

$$\sqrt{54} \cong \frac{\frac{54}{7.3} + 7.3}{2} = 7.349$$

This answer, when squared, yields 54.008, which represents better than 0.02% accuracy.

John A. Sego, K9DHD

active microphone impedance match

It seems that fate, often perverse, always arranges for those high-quality microphones that show up in surplus outlets to be of the low-impedance type. As nearly all amateur radio transmitters are designed for high-impedance microphone input, you have to use some sort of an impedance-matching device to mate the two. An additional complication appears in the situation of high-quality microphones often having an output that's not quite large enough to satisfy the average transmitter, even with an impedance matching (and voltage) step-up transformer.

Since I had just such a situation, plus a desire to rectify it, I built a transistorized unit that solved both elements of the problem quite satisfactorily. The solution had the additional positive attraction of costing nothing. Nothing, that is, if you have purchased a few surplus (and well-loaded) PC boards. That was the source of the parts I used, and the circuit boards had been acquired long ago just for a few specific parts, with the excess going into my junkbox. (Perhaps I should

be honest and state that *my* junkbox occupies an attic, an ex-garage and two backyard storage buildings!)

The transistor is an ideal impedance-matching device. In either the common-base or the common-emitter configuration it presents a low-impedance input circuit that will provide a reasonable impedance match for a 25-ohm microphone. Microphones, fortunately, are

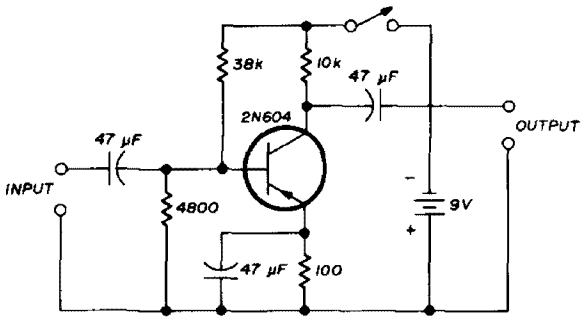


fig. 2. One-transistor microphone impedance-matching circuit matches low-impedance microphones to high-impedance inputs common with amateur ssb transceivers.

very tolerant of match. Just don't work them into too low a load impedance.

For my project, I selected the common-emitter configuration. The selection was not based upon any overweening factor; the common-base probably would have been just as satisfactory. Much the same attitude prevailed in the selection of component values. There is a very wide latitude of choice, within which the exercise of reasonable judgement determines what you'll use. The values given in the schematic diagram are ones that work and work well, being mutually compatible for desired gain and frequency range as well as low waveform distortion.

A PNP transistor was used, picked because of the high h_{fe} of the 2N604 lifted off a surplus circuit board. It must have been a good choice, for gain is more than ample. The output is appreciably greater than that from a transformer designed expressly for microphone impedance matching.

I built my unit on a piece of insulating board of uncertain origin and composition. Holes were drilled for flea clips and for the three leads of the transistor. These leads were bent over and soldered to adjacent flea clips. All components were mounted on flea clips, with the physical arrangement of parts almost exactly following the schematic diagram shown in **fig. 2.**

The completed unit enabled an Electro-Voice Model 637 microphone to provide all the audio input needed by my Drake TR-4 transceiver. On-the-air checks of audio quality have been quite pleasing, a compliment to both the microphone and the impedance-match/amplifier.

Carl Drumeller, W5JJ

using the Shure 401A microphone with the Drake TR-4

The audio passband of the Drake TR4 transceiver, when used with the recommended Electrovoice microphone, has insufficient gain if the operator has a deep, bass voice. This can be improved by using a Shure 401A microphone, but the original microphone circuit of the TR4, consisting of 1000-pF blocking capacitor and 6.8 megohm grid leak, is unsuitable for the Shure 401A.

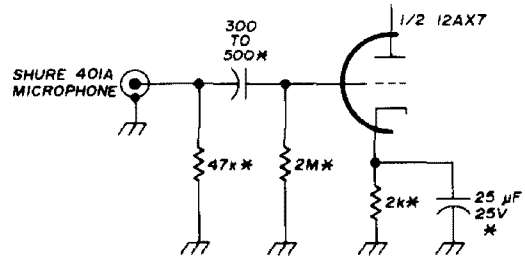


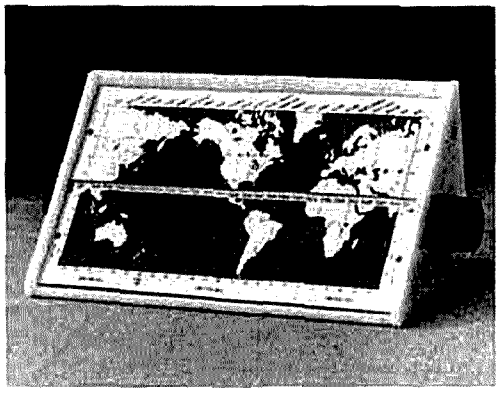
fig. 3. Modifications to the TR4 speech circuit for use with the Shure 401A microphone. This microphone is particularly suitable for operators with deep, bass voices.

The solution to this problem is shown in **fig. 3.** The new components are indicated by asterisks. In my rig the new components are clustered around the tube base. The existing printed-circuit board assembly is left intact except for the connection to the grid of the 12AX7.

Harry Manning, G3XOM

new products

graphic time-teller



An operating aid that will interest hams active in DX work, as well as foreign broadcast listeners, has been introduced by U-J Industries. By the turn of a knob the operator may position the drum dial, viewed through an opening in the map panel, so that his local time appears at the map line for his time zone. The standard time for all other time zones may then be read from the dial at those zone lines, and the GMT hours are shown in each margin of the map. If on daylight saving time, the operator simply positions his local daylight time at the meridian line to the east of his actual zone line. World standard times and GMT hours remain correct.

Schedule planning is greatly facilitated by the quickness with which the operator may turn to other times for his location for a look at other-country times at those other hours. This is a flexibility not realized through the use of tables or 24-hour clocks having countries listed around the dial.

The map panel features enamel on aluminum, with the ocean areas in blue and other features in black. In the top margin is shown, for each time zone, the name of one or two countries (or cities or other areas) that are within the zone. In the other margins latitude and longitude figures are shown. To add to the visual presentation, the hours of daylight are printed in red and the hours of darkness in black. This is an aid in determining conditions along the signal path. The cabinet is hand-crafted of acrylic sheet, and is slightly over nine inches long.

List price of the *Time-Teller* as illustrated is $18.50 plus postage. A complete kit is available at $14.75, and a basic kit is $11.75. The latter includes map panels, assembled drum dial with shaft and 2-color chart, angles and other hardware, but not cabinet (cabinet drawings are included) or knobs. Additional information is available from U-J Industries, 6605 Shoal Creek Boulevard, Austin, Texas 78757, or use *check-off* on page 110.

eye-ball QSL cards

These new eye-ball QSL cards are designed for the radio amateur who actively participates in hamfests, club meetings or eye-ball QSOs. Eye-Ball QSL Cards are simulated engraved business cards printed on deluxe citation card stock with a choice of three colors: black, blue and red. ARRL or club emblem cuts may also be included in your layout. There is no need to give away expensive QSL cards when you can give an Eye-Ball QSL Card for less than 1 cent each.

These eye-ball QSL cards are available at $7.95 per thousand for one color or $8.95 per thousand for two colors, postpaid (add $1.00 for ARRL or club emblem cuts). Available from Lecours Enterprises (KØOJW), 814 Riderwood Drive, Hazelwood, Missouri 63042.

More Details? CHECK–OFF Page 110

antenna noise bridge



Amateur radio antennas, including two-meter fm systems, can now be tested for resonant frequency and impedance with the Omega-t Systems Antenna Noise Bridge. The Antenna Noise Bridge, models TE7-01 and TE7-02, combine precision and measurement flexibility normally found only in a collection of expensive laboratory equipment.

Able to be used over the entire range from 1 to 300 MHz, the antenna noise bridge allows testing and tuning for optimum performance from an antenna and receiver without the use of additional equipment. It also replaces any other testing equipment, specifically vswr bridges, with a more accurate test system.

The antenna noise bridge can be used to test beams, whips, dipoles, quads or complete antenna systems by connecting the instrument between the antenna to be tested and the station receiver. For more information, write to Omega-t Systems, Inc., Box 1010, Richardson, Texas 75080, or use *check-off* on page 110.

veroboard kit

Vero has introduced their new Veroboard Kit DIP 24 for use with integrated circuits. Engineers, technicians and experimenters will enjoy using the DIP 24 Kit as it eliminates the need for etching, drilling and making terminals. The boards are used for mounting and interconnecting dual-in-line IC packages for development applications, as well as for production runs where variations of the basic design may be required.

Supply lines are positioned adjacent to all DIP locations and are connected to the

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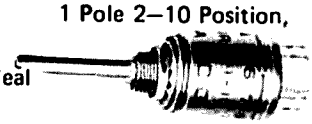
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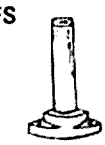
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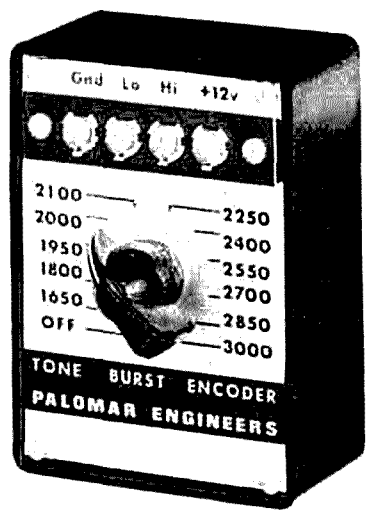
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outer edge contacts. The layout pattern permits the mounting of packages having any number of terminations provided they are on 0.1-inch centers. When fitted in horizontally mounted card frames the DIPs are positioned in parallel vertical rows. The boards are punched with a matrix of holes on 0.1" centers ready to accept dual-in-line packages or IC sockets. Plain holes off the copper tracks can be used for terminal pins.

The Vero DIP Kit consists of one 11824 board, which is 4.5 x 6.5 inches, with 22 contacts on two sides, a 44-pin edge connector, card extraction handle, design sheet and 10 14-pin DIP sockets. The DIP 24 Kit sells for $16.22. For more information, write to Vero Electronics Incorporated, 171 Bridge Road, Hauppauge, New York 11787, or use *check-off* on page 110.

tone burst encoder



Palomar Engineers has announced a new tone burst encoder for use with two-meter fm repeaters and in other remote-control applications. Using LC tuned circuits and an IC amplifier it features accurate tone frequency, low drift, and immunity to rf pickup and interference.

An important application is tone-burst entry to two-meter repeaters. Many repeaters require a short burst of an audio

More Details? CHECK–OFF Page 110

tone for turn-on. Each repeater in a local area and on the same or nearby frequency uses a different tone; this prevents accidental turn-on of other repeaters. Eleven tones, from 1650 to 3000 Hz, are in widespread use. The Palomar Engineers encoder produces all eleven tones as selected by a front-panel switch.

The Palomar encoder has a built-in half-second timer. The time is adjustable and can be wired to produce a continuous tone for other remote control applications.

The Encoder is housed in a compact 4x3x2-inch case with an anodized aluminum control panel. The required +12 Vdc power is normally taken from the associated transceiver. Outputs for both high and low impedance microphones are provided so the unit may be used with all the popular fm transceivers. Price is $37.50, postpaid. For a free descriptive brochure, write to Palomar Engineers, Box 455, Escondido, California 92025, or use *check-off* on page 110.

fast-scan/slow-scan tv camera

The Thomas Electronics and Engineering Company has recently announced the availability of their new, all solid-state, fast-scan/slow-scan television camera, the model HCV-1B. This new camera features a built-in ac power supply and rf output on fast scan for viewing on a standard tv set. There is a provision for the addition of an audio sub-carrier module for home entertainment and recording of fast-scan signals on video tape.

Slow-scan output meets FCC standards with white at 2300 Hz, black at 1500 Hz and sync at 1200 Hz. Amplitude is adjustable, zero to 3 volts p-p. Output impedance is 1000 ohms. The output spectrum also meets published Bell System tariffs for voice couplers.

The video output on fast-scan is 1.5 volts p-p. The rf output is 1.5 millivolts p-p, adjustable for channels 2 through 6. With this camera fast-scan video or rf may be monitored at the same time slow-scan is being televised.

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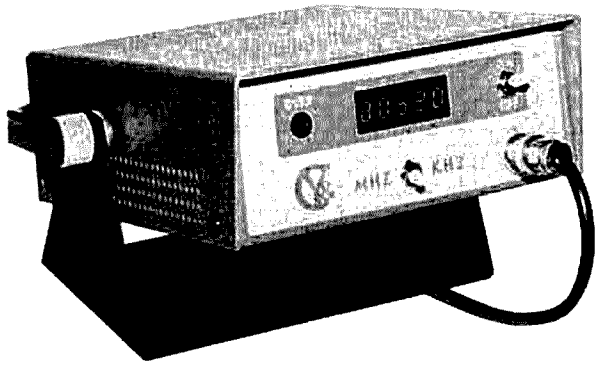
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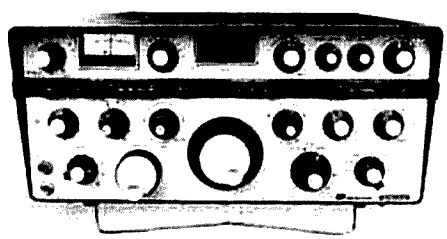


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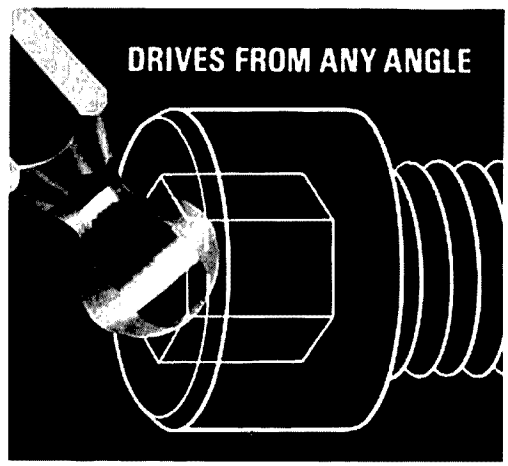
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Price, ready to operate with a Soligor 25mm, f1.9 lens or Cosmicar 2519 is $395.00. Price less lens is $365.00. For more information, write to Thomas Electronics and Engineering Company, Inc., Post Office Box 572, Hendersonville, Tennessee 37075, or use *check-off* on page 110.

unique hex-socket drivers



Xcelite is now manufacturing a line of Allen hex-type screwdrivers and interchangeable blades with an unusual "ballpoint" tip design that achieves a speed and ease in engaging and turning that is unattainable with conventional drivers.

The tools work at any angle, thus being able to handle hex socket screws which, because of obstructions, cannot be reached straight-on. Because they slip into sockets more easily and faster than regular hex socket drivers, they simplify adjustments and speed up work, even in normal situations.

Nine sizes, from .050 through 3/16 inch, are available; fixed handle types,

More Details? CHECK–OFF Page 110

singly or a complete set in a handy roll kit with extra pockets for associated tools are available, as are interchangeable Series 99 blades singly or in a compact set that includes a regular detachable handle, extension, and a transparent plastic case. Complete information, specifications, and prices are contained in new Bulletin/Price List No. 273L which may be obtained by writing Xcelite Incorporated, Orchard Park, New York 14127, or by using *check-off* on page 110.

fm and repeaters for the radio amateur

The ARRL has recently introduced the latest in their long list of books for the radio amateur. This new book, "FM and Repeaters for the Radio Amateur" represents several years work on the part of the editor, Thomas McMullen, W1SL, and many members of the league staff as well as contributions from amateurs from all over the states. Although some of the material in this book was originally published in *QST*, the majority is new, including many new projects and circuits that haven't appeared in print before.

Chapters cover such topics as fm receivers, fm transmitters, mobile and portable equipment, antennas for base and mobile stations, repeaters, repeater controls and accessories, repeater technical problems and cures, testing fm gear, using surplus fm equipment, tips on buying used fm gear, repeater club organization, fm operating practices, suggestions for completing FCC repeater applications and the latest FCC regulations that pertain to repeater use.

For the amateur who is interested in vhf fm, this book is an essential addition to his library. If you are having a problem with your repeater system, chances are that you will find the answer in this excellent manual. Soft-bound, 232 pages, hundreds of photographs and drawings, $3.00 from your local dealer or from Comtec Books, Greenville, New Hampshire 03048.

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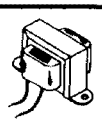
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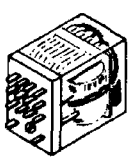
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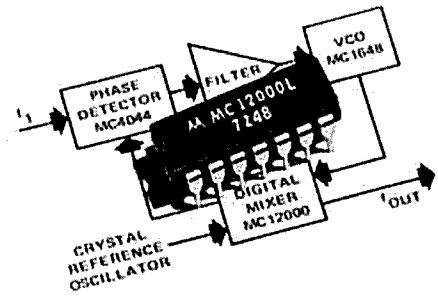
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digital mixer for phase-locked loops



Motorola Semiconductor Products has just announced introduction of a digital mixer for use in phase-locked loops (PLLs). Numbered the MC12000, the digital mixer generates an output frequency which is the difference between its two input frequencies.

The mixer consists of a D-type flip-flop, together with TTL to ECL and ECL to TTL translators. Using the MC12000 in a phase-locked loop, frequencies (one or more) up to 250 MHz can be generated economically and without the need for tuned circuits. The output frequency may be either a single fixed frequency or a series of programmable frequencies (using additional circuitry).

Phase-locked loops are electronic servo loops which cause an oscillator to follow exactly the phase of a reference signal. Either an analog or a digital approach may be taken toward the design of a phase-locked loop. The analog method is useful for *signal detection* (as in fm demodulation), while the digital approach is best suited for *signal generation* (as in multi-channel frequency synthesis and computer timing chains).

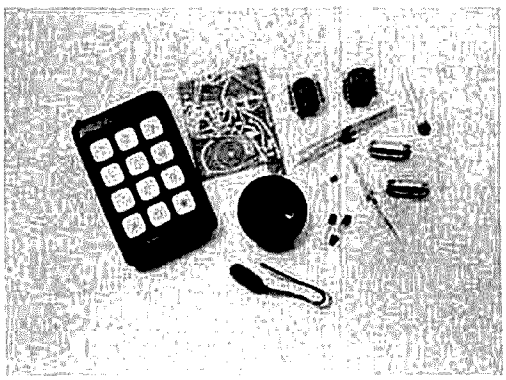
It is the latter application area, signal generation, for which the new Motorola digital circuit is intended. The mixer is expected to find use as a prescaler in phase-locked loop applications where the voltage-controlled oscillator operates above 10 MHz and a relatively narrow tuning range is required.

The MC12000 is available off-the-shelf in a 14-pin ceramic dual in-line package. Pricing in small quantities is $7.50. For

More Details? CHECK–OFF Page 110

more information, contact the Technical Information Center, Motorola Inc., Semiconductor Products Division, Box 20924, Phoenix, Arizona 85036, or use *check-off* on page 110.

touch-tone generator

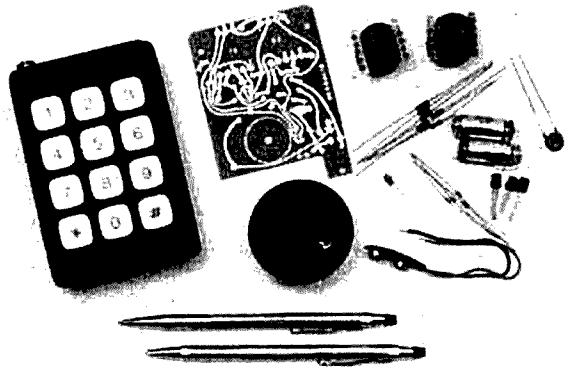


Interface Technology has announced a small Touch-Tone generator for repeater keying or dialing through on a phone patch. The unit generates the standard 12 frequency pairs used for Touch-Tone dialing by the telephone companies. It is designed so that the output can be used in several ways. For one, a speaker included in the kit can be mounted internally and the unit simply held up to the microphone. The speaker can also be mounted in a small, remote case (optional) and connected by a cable to the generator; this allows the user to position the generator on a table or desk while the speaker is held up to the microphone. A third approach is to wire the unit directly into the microphone circuit, eliminating the speaker altogether.

There are no switches or controls on the unit other than the key pad, so no current is drawn from the standard 9-volt transistor battery until the operator touches a key to generate a tone. This feature insures long battery life and simplicity of use.

The unit is packaged in an attractive black molded plastic case and weighs only 8 ounces. It sells for $33.95 in kit form or $53.95 assembled and tested. Write to Interface Technology, Inc., Post Office Box 24565, St. Louis, Missouri 63141 for more information, or use *check-off* on page 110.

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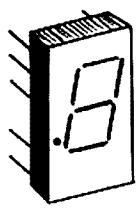
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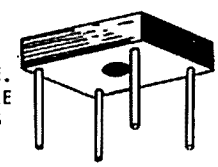
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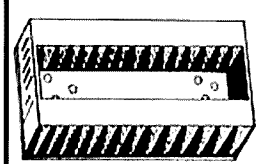


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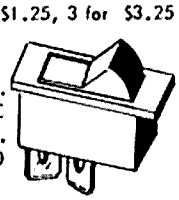
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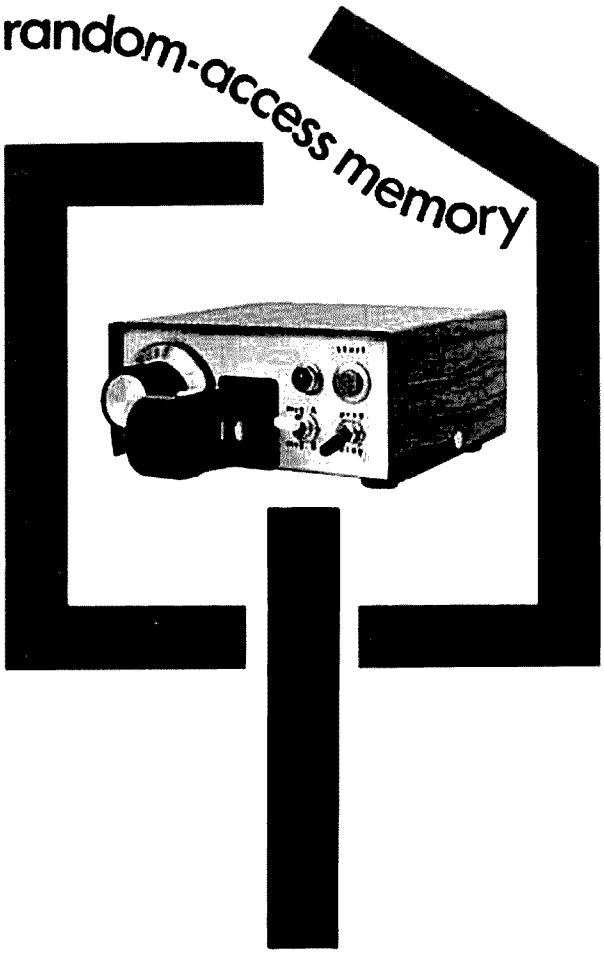
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ham radio magazine

OCTOBER 1973

low-cost electronic keyer with random-access memory



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October, 1973
volume 6, number 10

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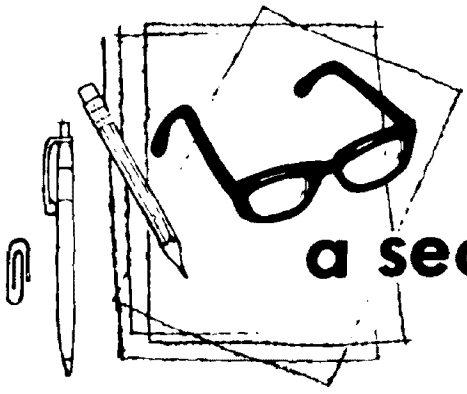
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a second look

by jim fisk

The editor of a technically oriented magazine such as *ham radio* must wear several different and diverse hats. In fact, I could use up this entire page describing all the details that need attention to keep the magazine running smoothly. However, I'd like to talk for a moment about one very important editorial task that instills confidence in the reader, and one that doesn't. This task, which I share with the rest of the editorial staff, is that of researcher and seeker of truth.

Most of the articles published in *ham radio* are contributed by readers who want to share an idea or the details of a particularly successful project. Authors range from enthusiastic hams who have never written anything more than a short story for their English professor to fellows with engineering backgrounds who make their livings in front of a typewriter. All want to share an idea and I welcome the output of anyone who is interested in contributing something that will benefit all hams.

Budding authors often ask, "What kind of articles are you looking for?"

That's a difficult question to answer because many new manuscripts come across my desk every day, but generally speaking, I am looking for simple construction projects that the average reader can complete in one or two weekends. Larger projects are also welcome, but most *ham radio* readers must split their spare time between amateur radio and other interests, so they don't have time to build Chinese copies of complex electronics equipment.

Once a month I set one or two days aside to go over all manuscripts that have come in during the previous month. Since I seldom use more than a dozen articles in any issue, I don't accept more than that during any one-month period. This is sometimes a nearly hopeless task since there may be three-dozen or more manuscripts to be considered. The first things I look for are originality and interest value. If the contribution passes this test, the next thing I look for is technical accuracy and attention to detail.

The contributed article doesn't have to be a literary masterpiece to be accepted. If you have a good idea and it's well documented, if the illustrations and technical discussion are clear and accurate — you may have a winner! On the other hand, if the article rambles from one topic to another, covers ground that has been over thousands of times before, or presents inaccurate or misleading information, you will receive a rejection slip.

If your article has been accepted for publication, don't expect to see it published in the very next issue. The production times for a monthly magazine are probably much longer than you ever imagined. The articles for this issue, for example, were being prepared for publication during the month of June. As you are reading this we are putting together the material for the February, 1974, issue of *ham radio.*

Incidentally, my staff and I are fairly adept at ferreting out technical inaccuracies, but despite research and keeping the mailman busy between our editorial office and the author, errors do occasionally creep into the magazine. Contributing authors can help by carefully checking out their facts before submitting the article. Errors can cause considerable misery to the builder, and as publishers *we* are taken to task for the error. So, before you send in that next article, spend some extra time going over the text and illustrations — it might save some later embarrassment.

Jim Fisk, W1DTY
editor



low-cost electronic keyer with random-access memory

Circuit details of an electronic keyer with a 512-bit random-access memory that costs less than $40

Michael J. Gordon, Jr., WB9FHC, 1615 Elmwood, Wilmette, Illinois 60091

Although many circuits for electronic keyers with a memory have appeared in the literature, recent developments in the IC industry have rendered these designs obsolete. These developments include the introduction of MOS random-access memories which are compatible with TTL ICs. It is now possible to build an electronic keyer with 512 bits of storage for a total parts cost of $40.00.* Thanks to the miniature MSI and LSI devices, there are only eight ICs in the unit, which fits on a 2-1/2- by 4-inch circuit board, power supply included!

circuit

The circuit is not very complex and can be divided into two parts: the keyer, which automatically produces dots and dashes, and the memory section and its associated logic.

The keyer circuit is fairly standard and is designed for use with a standard paddle. Speed is variable from roughly 6 to 60 words per minute; the characters are

*A complete kit of parts, including circuit board, is available for $40.00 from Psynexus Systems, P.O. Box 277, Glencoe, Illinois 60022. The drilled glass-epoxy circuit board is available for $4.00. All items post-paid.

self-completing. Details of how this portion of the circuit operates are best understood by looking at the timing diagram, fig. 1, and by examining the logic diagram, fig. 2. Output from the keyer circuit is taken at two points, labeled DATA and $\overline{\text{DATA}}$ on the logic diagram. The DATA line goes directly to the output gate, where it is inverted and fed to the output keying transistor, a pnp device with a breakdown voltage of 150 volts. The $\overline{\text{DATA}}$ line is fed to the memory inputs.

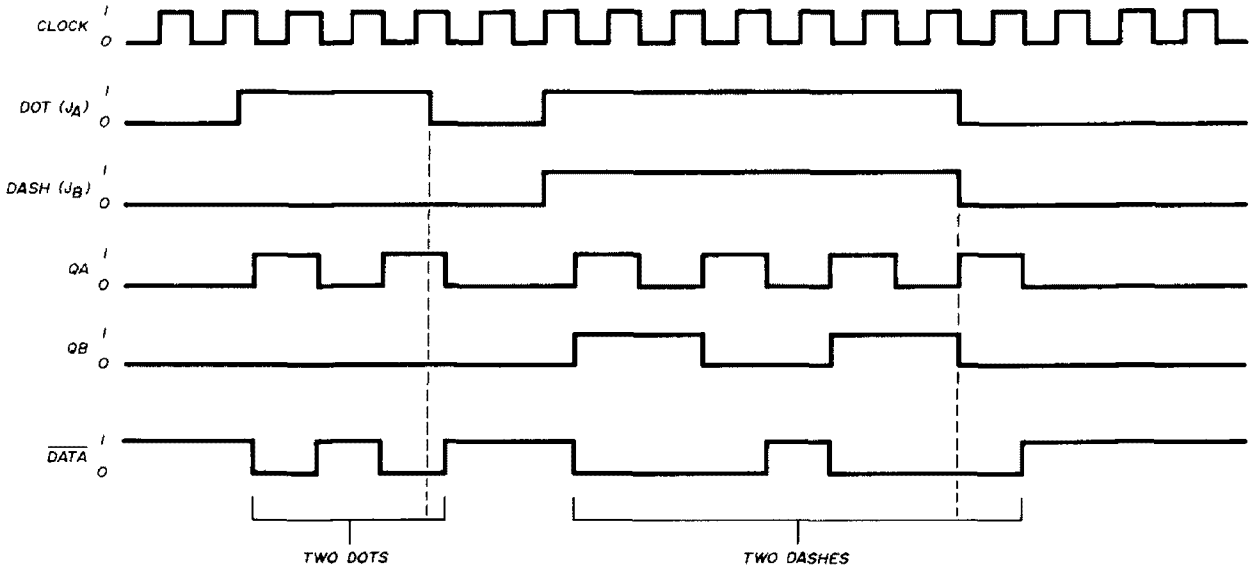


fig. 1. Timing diagram for the electronic keyer. Note that closing the dash contacts puts a logic 1 on the J pins of both flip-flops A and B (fig. 2). Characters are self-completing.

memory

The two memory ICs used in this design are Signetics 25L01B 256-bit random-access memories. Each is an MOS LSI circuit containing thousands of transistors on a single chip. The particular memory cell that is being readout at a given time is determined by the status of the eight address lines. A chip-select control is provided; when a chip is not selected the input buffers are disabled and the output buffers are cut off, effectively taking the device out of the circuit. This allows the two ICs to be connected in parallel, simplifying both circuit board design and the logic requirements. It is this chip-select feature that makes the random-access memory more

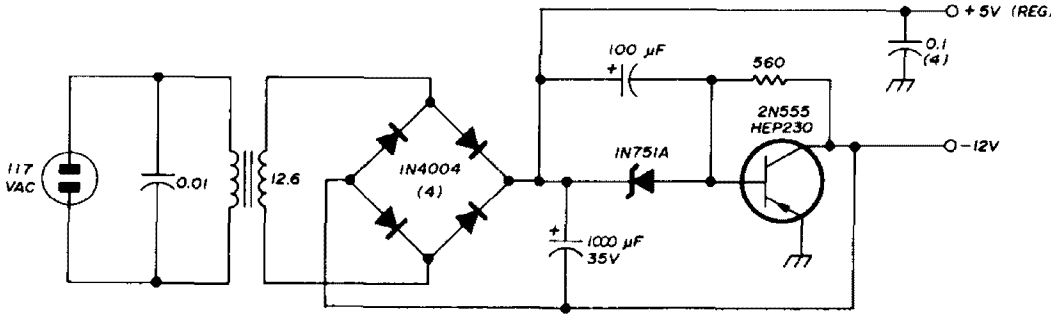
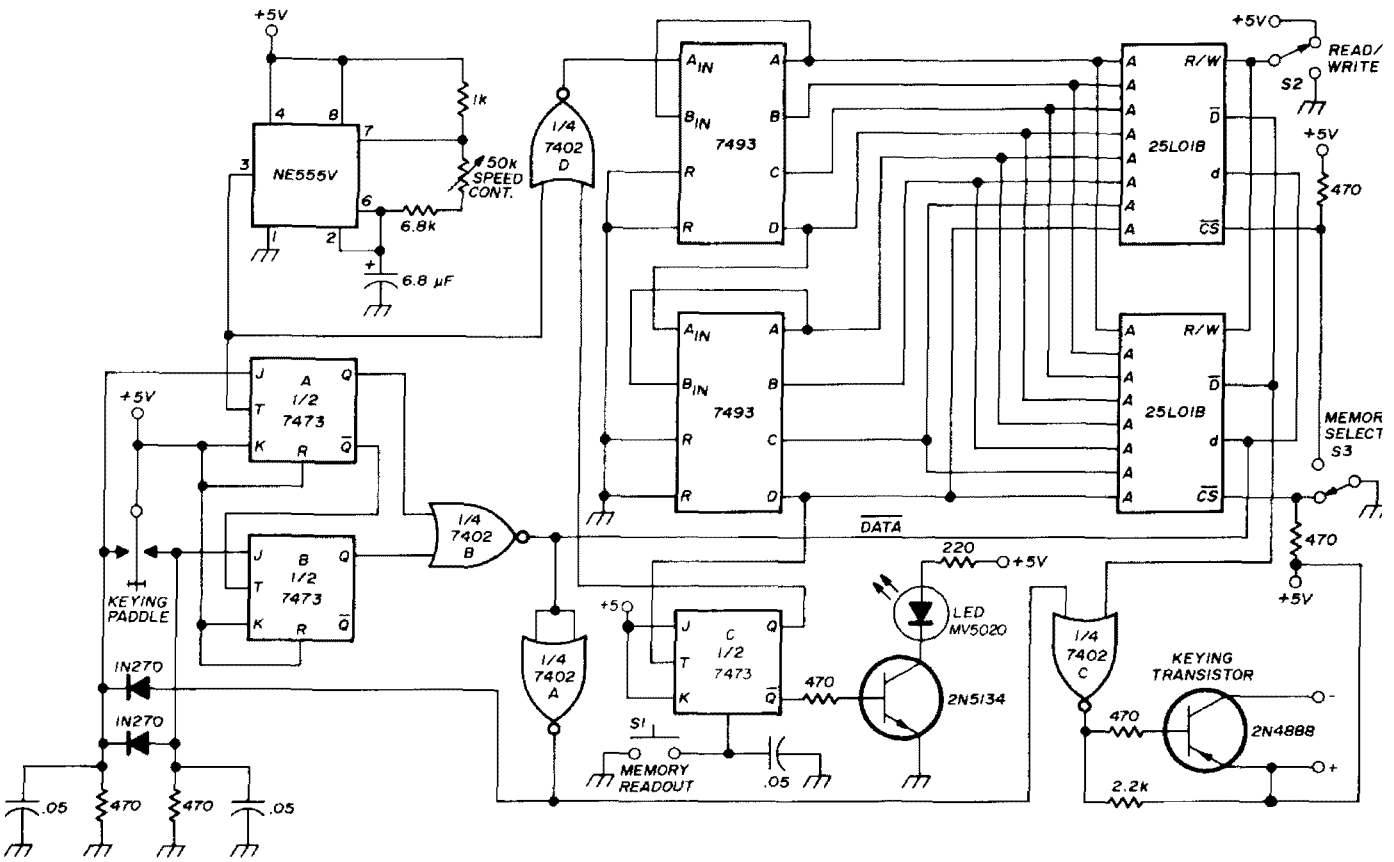


fig. 3. Power supply for the electronic keyer. The negative 12-volt line powers the MOS memory circuits.



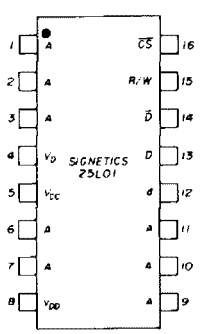


fig. 2. Logic diagram of the low-cost electronic keyer with 512 bits of memory. Memory is provided by two Signetics 25LO1B read-only memory ICs. Breakdown voltage of the 2N4888 keying transistor is 150 volts.

cost-effective than the shift register usually used in this type of circuit.

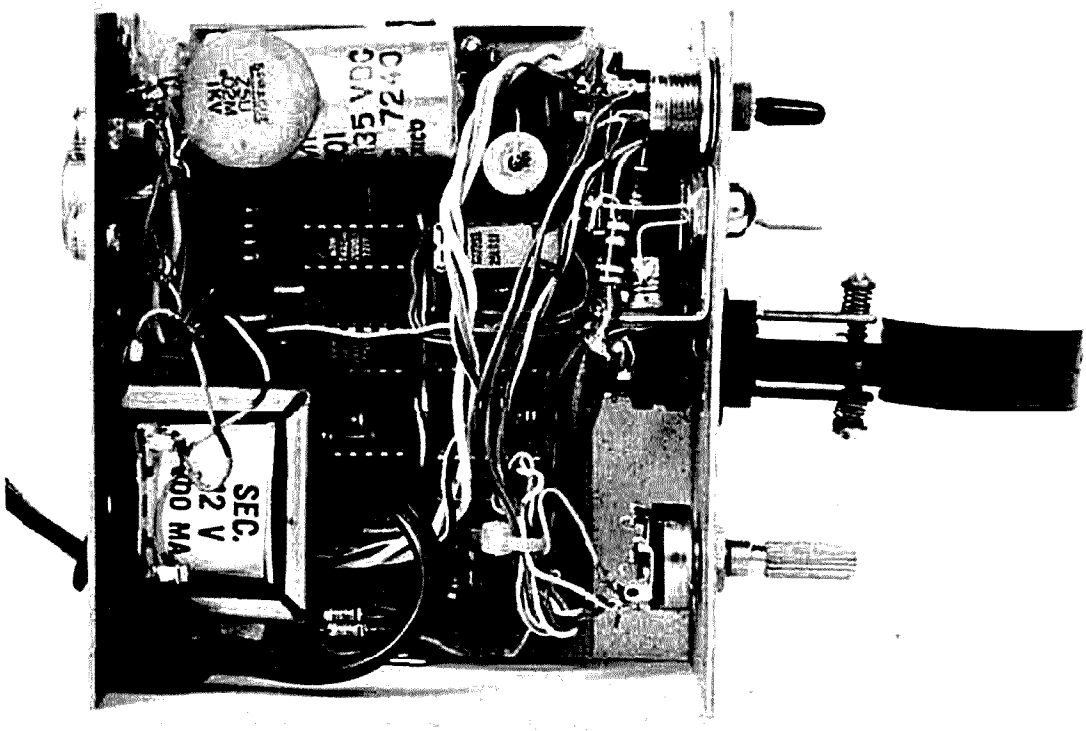
The MSI circuits, both four-bit binary counters, are used to cycle the memory through its 256 bits during the read and write cycle. Normally, the Q output of flip-flop C is in the logic 1 state, and the output of gate D is held low. However, when S1 is closed flip-flop C is reset, Q goes to zero and clock pulses are fed to the memory address counters which cause each of the 256 locations in the selected memory to be accessed.

On the 256th clock pulse, the last stage in the counter triggers flip-flop C — Q once again is returned to the logic 1 state and a memory cycle is complete. During the cycle, Q is in the logic 1 state and the LED is lighted, indicating a memory cycle in progress. If the read/write switch, S2, is in the read position during a cycle, the data in the memory is sent to the output gate and thus keys the transmitter. If this switch is in the write mode, whatever is sent by the paddle is recorded by the selected memory as well as sent to the output gate. Switch S3 selects one memory of the two available.

Clock pulses for the entire circuit are generated by a Signetics NE555V timer IC. The speed is variable over an extremely wide range by a single 50k pot. The NE555 is an excellent choice for this application because of its TTL compatible output, small size and low power consumption.

power supply

The power supply is simple, efficient and effective (see fig. 3). It consists of a *full-wave bridge feeding a discrete regulator*. The voltages necessary for the proper operation of this circuit are +5 volts and −12 volts. The most positive point in the



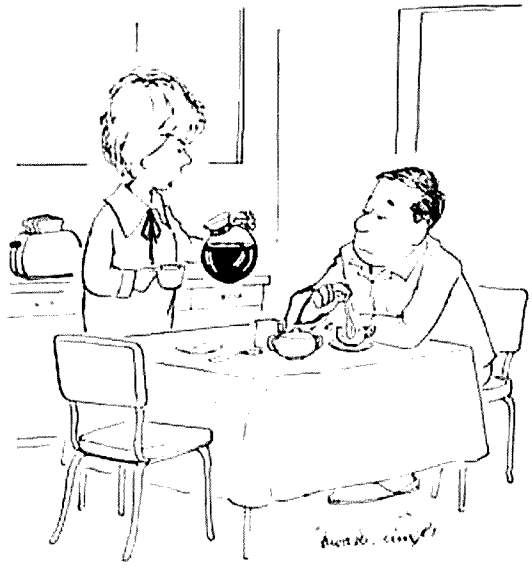
Construction of the electronic keyer with memory. Timing and memory components are mounted on circuit board. Power supply components are mounted on rear wall of the enclosure.

circuit is called +5, and the regulator keeps ground five volts negative with respect to this point. The most negative point, the other side of the filter capacitor, is roughly 12 volts below ground. Since the 12-volt supply is not really critical, it is not regulated. As long as it provides between −10 and −13 volts, the circuit should operate properly. Four 0.1-μF ceramic bypass capacitors are placed at various points on the circuit board between V_{cc} and ground to absorb current spikes generated by the TTL output logic.

operation

Completely assembled, this electronic keyer is a joy to operate. Programming is simply a matter of pressing the start button and sending the desired message. Readout is even simpler, requiring only the push of a button. There is no need to switch between the manual and automatic mode of operation, since this is done automatically by the logic. Now, you can have your own keyer with 512 bits of storage, at a cost that is a far cry from the $200 and up that commercial units command.

ham radio



"Look, 'Mr. Ham Operator,' if you want more coffee just ask for it, and stop tapping out —·— ——— ·—· ·—· · · on your cup."

continuous-phase audio-shift keyer for RTTY

Circuit details for an audio-frequency shift keyer which introduces no phase disruptions when switching from mark to space, or vice versa

Eric Kirchner, VE3CTP, Ontario Science Center, Don Mills, Ontario

The desirability for the absence of phase discontinuities in the output signal of audio frequency-shift RTTY keyers has been stressed in previous articles.[1,2] These phase discontinuities appear at that point in time when the AFSK generator frequency is shifted from mark to space, or vice versa. The switch-over from one frequency to the other appears as a disruption along the sinusoidal waveform, as shown in fig. 1A.

A phase disruption such as this causes over- and under-shoots which manifest themselves as fast amplitude changes of the transmitted rf envelope. This leads to

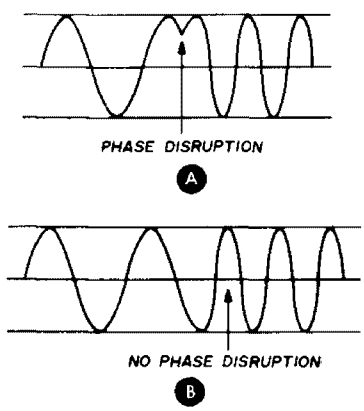


fig. 1. In many AFSK generators severe phase discontinuities are introduced when switching from mark to space, or vice versa, as shown in A. This results in undesirable clicks, similar to CW key clicks. With the AFSK generator circuit described here there is no phase discontinuity when shifting frequencies, B.

clicks, similar to CW clicks, which can be heard on either side of the RTTY signal. Needless to say, these clicks interfere with stations operating on adjacent frequencies.

Although AFSK generator circuits which eliminate these phase discontinuities have been described in the past, they were complex and expensive, and used toroid inductors. The circuit described here was designed with simplicity and state-of-the-art in mind, and uses no toroid inductors.

circuit

The circuit for the continuous-phase AFSK generator is shown in fig. 2. Integrated circuits U1 and U2 constitute the audio-frequency oscillator. The output of U2, a National Semiconductor LM311H, is a square wave which is fed through R4 to the input of U1, a National Semiconductor LM301AN. U1 operates as an active filter whose frequency is determined by C1, C2, R1 and R2. The output of the active filter is capacitively coupled to the input of U2. The loop is closed and oscillation occurs.

The sine-wave output is available at pin 6 of U1. This sine wave crosses the zero voltage point at precisely the same time the square wave at pin 7 of U2 changes polarity as shown in fig. 3. This square-wave transition can therefore be used to command the switchover from one audio frequency to the other to occur *only* at the zero-voltage crossover of the sine wave. This will provide a phase-continuous output.

The switchover command is accomplished in the following way. The square-

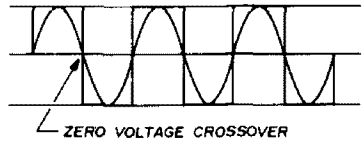


fig. 3. In the circuit of fig. 2 the zero-voltage crossover of the sine-wave output coincides with the zero-voltage crossover of the square wave. The square-wave transition is used to control the precise moment mark-space switching occurs.

wave output from U2 is fed to transistor Q2 which operates as a voltage-level changer. The output swing from U2 is about ±10 volts, while the maximum input requirement for U3, a TTL master-slave J-K flip-flop, is from zero to +5 volts. The square-wave signal at the collector of Q2 toggles the input of U3. When the keyboard contacts connected to the input of gate U4 are opened, the polarity of the voltage at pins 6 and 8 of U4 invert, and the Q output of U3 will change its state then, and only then. This

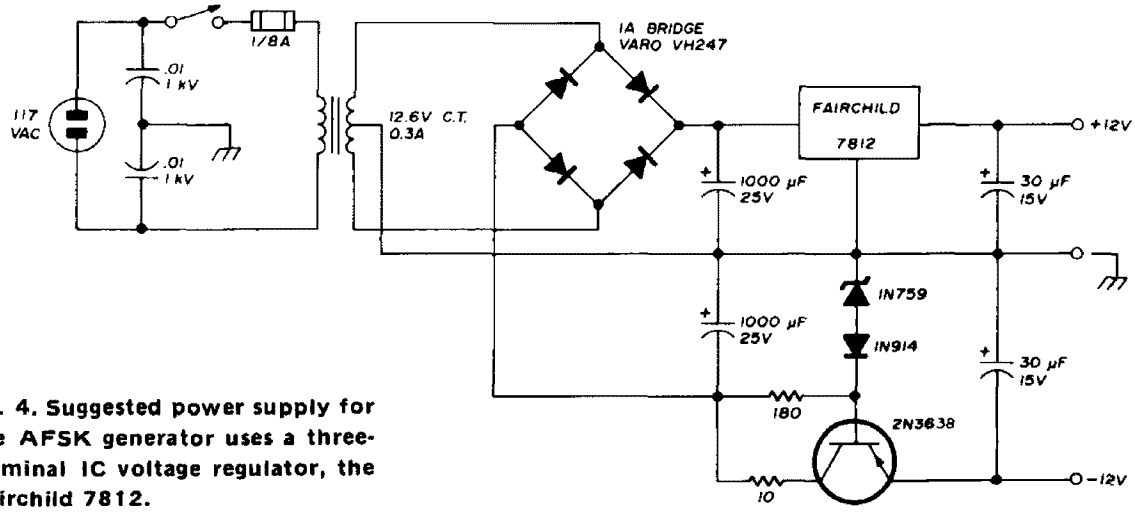


fig. 4. Suggested power supply for the AFSK generator uses a three-terminal IC voltage regulator, the Fairchild 7812.

only happens when U3 is toggled by the negative-going transition of the square wave at Q2, which occurs at the zero-voltage crossover of the sine wave. At that instant the Q output of pin 8 of U3 goes low and brings transistor Q1 into conduction.

Transistor Q1 operates as a switch, effectively paralleling the resistance network R1 and R6 with an additional resistor, increasing the frequency of the audio-frequency generator. When the keyboard contacts are closed again, the same sequence occurs in reverse, switching back to the lower frequency at precisely the zero-voltage crossover point of the sine wave. In this way phase discontinuities are avoided and the AFSK sine wave is sufficiently pure for use with ssb transmitters.

The audio frequencies I chose were dictated by a desire to eliminate the need for a special carrier crystal in my ssb exciter. The lowest audio frequency of this keyer is at 1450 Hz, high enough to place the second harmonic at 2900 Hz, out of the passband of most modern amateur ssb transmitters. The highest audio frequency, at 2300 Hz, lies within the passband of this same equipment.

The AFSK generator requires +12 volts at 70 mA and -12 volts at 15 mA. A suitable power supply circuit is shown in fig. 4.

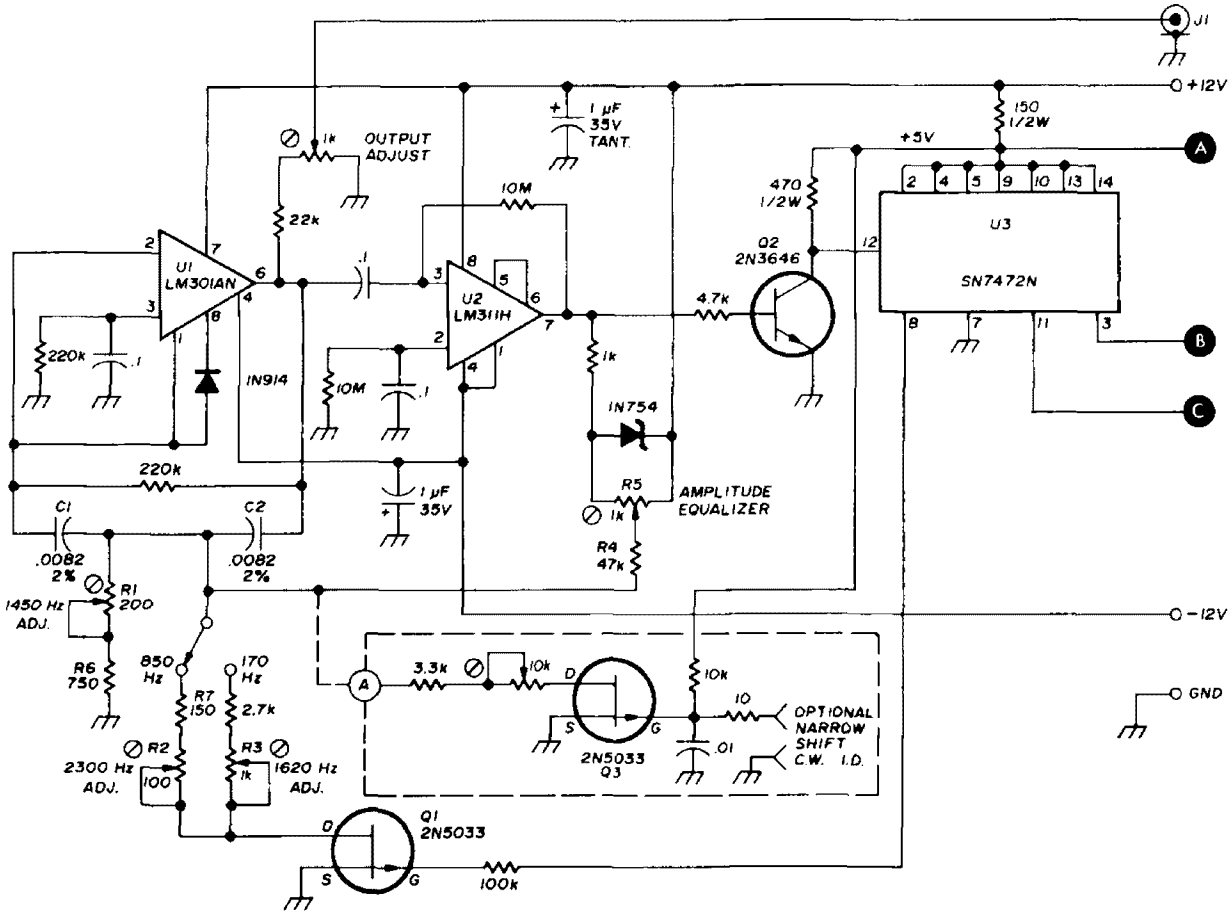


fig. 2. Circuit for the continuous-phase AFSK generator. Capacitors C1 and C2 are polystyrene types. All ICs are manufactured by National Semiconductor.

alignment

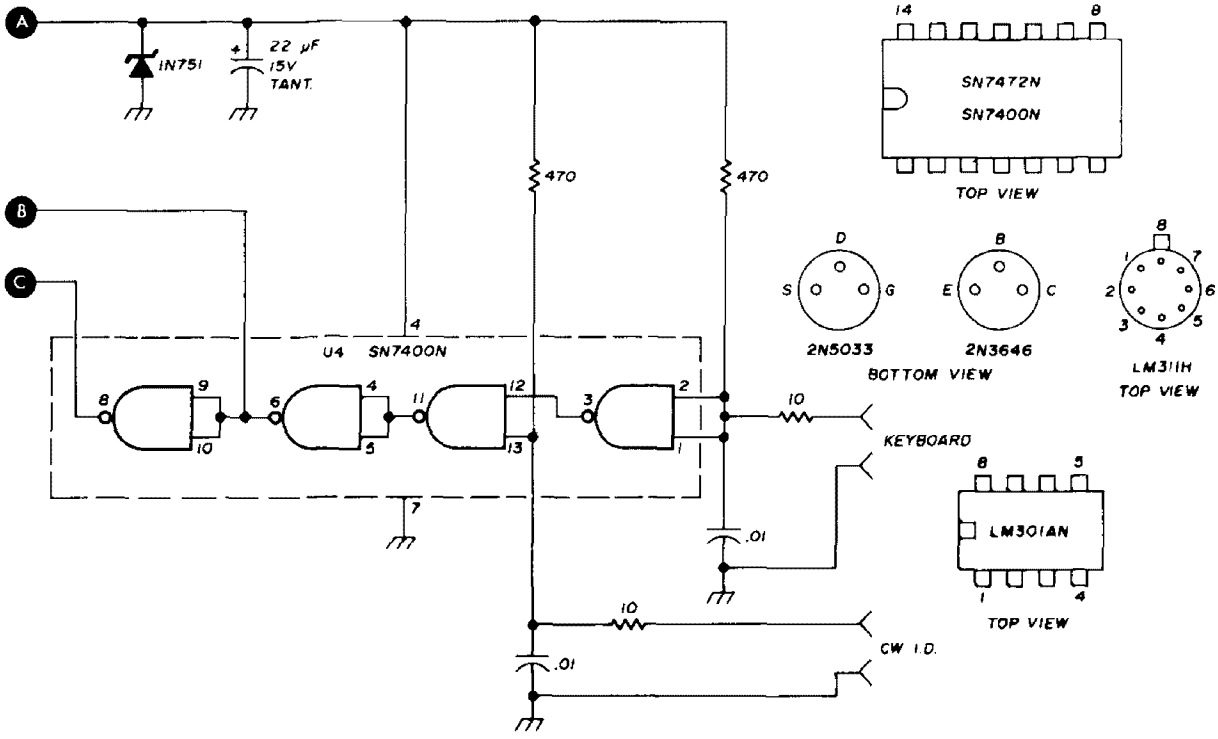
The best way to adjust the frequency and amplitude equalization of this AFSK generator is by using a digital frequency counter and an oscilloscope. These instruments are now owned by many amateurs so you should be able to enlist some help, if you don't personally own this test equipment. Once the alignment is completed it should hold for a long time.

First, connect the frequency counter

to the output jack, J1. Short the keyboard input terminals and adjust R1 for 1450 Hz on the counter. Set the switch to 850-Hz shift and adjust R2 for 2300 Hz. Set the switch to 170-Hz shift and adjust R3 for 1620 Hz. For easier adjustment you may want to install more expensive multi-turn trimming potentiometers at R1, R2 and R3.

optional circuit shown within the dashed lines in fig. 3. Simply connect point A of the optional CW identification circuit to point A in the main AFSK generator circuit.

Although the input terminals could be connected directly to the keyboard contacts, in most RTTY station setups they are not. Since it is desirable to copy the



Now, disconnect the counter from the output jack and connect a scope probe to pin 6 of U1. Set the switch to 170-Hz shift and, while simultaneously opening and closing the keyboard input terminals, adjust R5 so that there is no amplitude difference shown on the scope. There may be a slight amplitude difference when switching to 850-Hz shift, but this should be on the order of 2 percent or less, which is not objectionable.

additional notes

The CW identification circuit, as shown, results in the same frequency shift as the selected RTTY shift, 170 or 850 Hz. If you want to use a different CW identification shift, you can add the

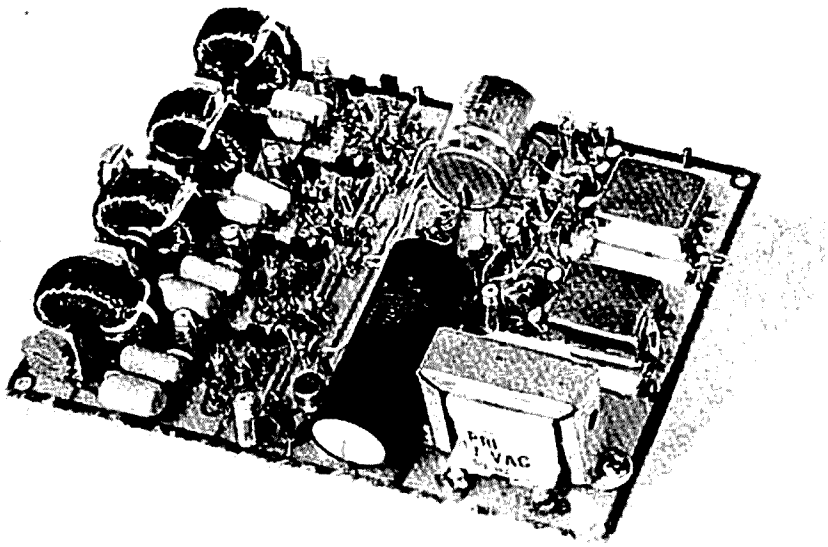
outgoing transmission on the printer, I have a mercury-wetted relay in series with the printer solenoid — the relay contacts are used to key the AFSK generator. A suitable relay for this purpose is the Potter and Brumfield JML-1061-81 with a 33-ohm resistor connected in parallel with the relay coil.

references

1. D.H. Phillips, W6FOO, "Radio Teletype using SSB Transceivers," *ham radio*, November, 1970, page 38.

2. D.H. Phillips, W6FOO, "Synchrophase AFSK Oscillator for RTTY," *ham radio*, December, 1970, page 30.

ham radio



multi-function touch-tone decoder

Complete construction details for a solid-state, multi-function Touch-Tone decoder using latching relays

Robert Heptig, KØPHF, Robert Shriner, WAØUZO, Pueblo, Colorado

With the advent of FCC Docket 18803, all repeaters are faced with the problem of shutdown control that is now defined by the FCC. The decoding system described here was designed to give absolute security in repeater control — it cannot be activated by anything other than a true Touch-Tone signal.

Two functions are utilized with latching-type relays. In the event of power failure the relays will remain in the position in which they were placed by the control operator. A time-out timer is also available in the circuit so that one of the functions will time out if desired.

For those readers who do not understand how Touch-Tone works, a short explanation seems to be in order. The Touch-Tone system was originally designed for dialing telephone services. It uses two tones for each digit, zero through 9 as well as * and #. These tones are generated by an oscillator in the

The lead photo shows a complete Touch-Tone decoder built by a Pueblo Ham, Jim Warner WBØ BTA, for use in muting his speaker when his wife is asleep. He's figuring on hooking up the second function to the coffee pot.

Touch-Tone pads. Fig. 1 lists the frequencies of these tones. Note that only 7 tones are used. The digit 1 is composed of 1209 HZ and 697 Hz, while the digit 2 is composed of 1336 Hz and 697 Hz, and so on, throughout the twelve digits which make up the Touch-Tone pad.

These tones are combined within the pad itself by various switch points. It is the job of the decoder to separate the various tone frequencies on receipt of the signal. If you don't thoroughly understand this go back and read it again, as you must understand how the ten Touch-Tone digits are made up of combinations of 7 tones. Also note that for explanation purposes, each tone is assigned a letter from A thru G. These letters that I have assigned to the various tones will be used for the balance of this discussion.

To use the Touch-Tone system, these tones must be decoded, changed into a dc voltage and combined in an AND gate which will form digits. Then, two of the digits are combined and used to close a relay. What you do with this relay closure is up to you. It may be used to shut down your repeater, disconnect the speaker of your base station, turn on your porch lights, start the coffee perking or whatever your imagination can conjure up.

There are many circuits that can be used to decode tones and cause a relay to open or close. One of the newest circuits is the phase-locked loop (PLL). This circuit was tried in many different configurations but its primary fault is that it will accept tones that are not on the exact tone frequency. Of course, this is an asset in some applications other than Touch-Tone decoders. The broad frequency response of the PLL can also cause problems if the repeater users use a Touch-Tone autopatch system — can you think of a better way to make the telephone company unhappy than to send a batch of off-frequency tones over their lines?

Other circuits using transistor decoders don't work too well as Touch-Tone decoders because of the inherent low impedance of a transistor, necessitating quite tight coupling of the tuned circuit which lowers the Q of the circuit.

The circuit shown in fig. 2 uses the high impedance of the field-effect transistor. This permits the use of very small coupling capacitors (.01 μF). The resulting Q of the circuit is sufficiently high to assure that the tones are on frequency. The high circuit Q also prevents accidental functions from occuring due to noise or excessive audio on the repeater.

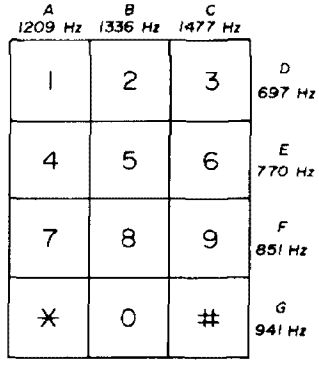


fig. 1. Twelve-button Touch-Tone pad showing tone frequencies and letter designations used for the purpose of this article.

the circuit

Now, let's get down to the nitty gritty of the circuit and see just what happens when the decoder sees a batch of frequencies coming in the front door. First of all, all frequencies are passed through potentiometer R1, which operates as a signal level control, and then fed to potentiometers R2, R3, R4 and R5. These 2.2-meg variables are also used as level controls, but only for the particular frequency of the circuit which they are in. When this multitude of frequencies reaches the 88-mH coil L1, and tuning capacitors C5 and C6 all except the selected frequency are passed straight to ground. The selected frequency is controlled by the tuning of L1, C5 and C6, which will be explained later. This frequency is tone A or 1209 Hz and is fed to the gate of Q1.

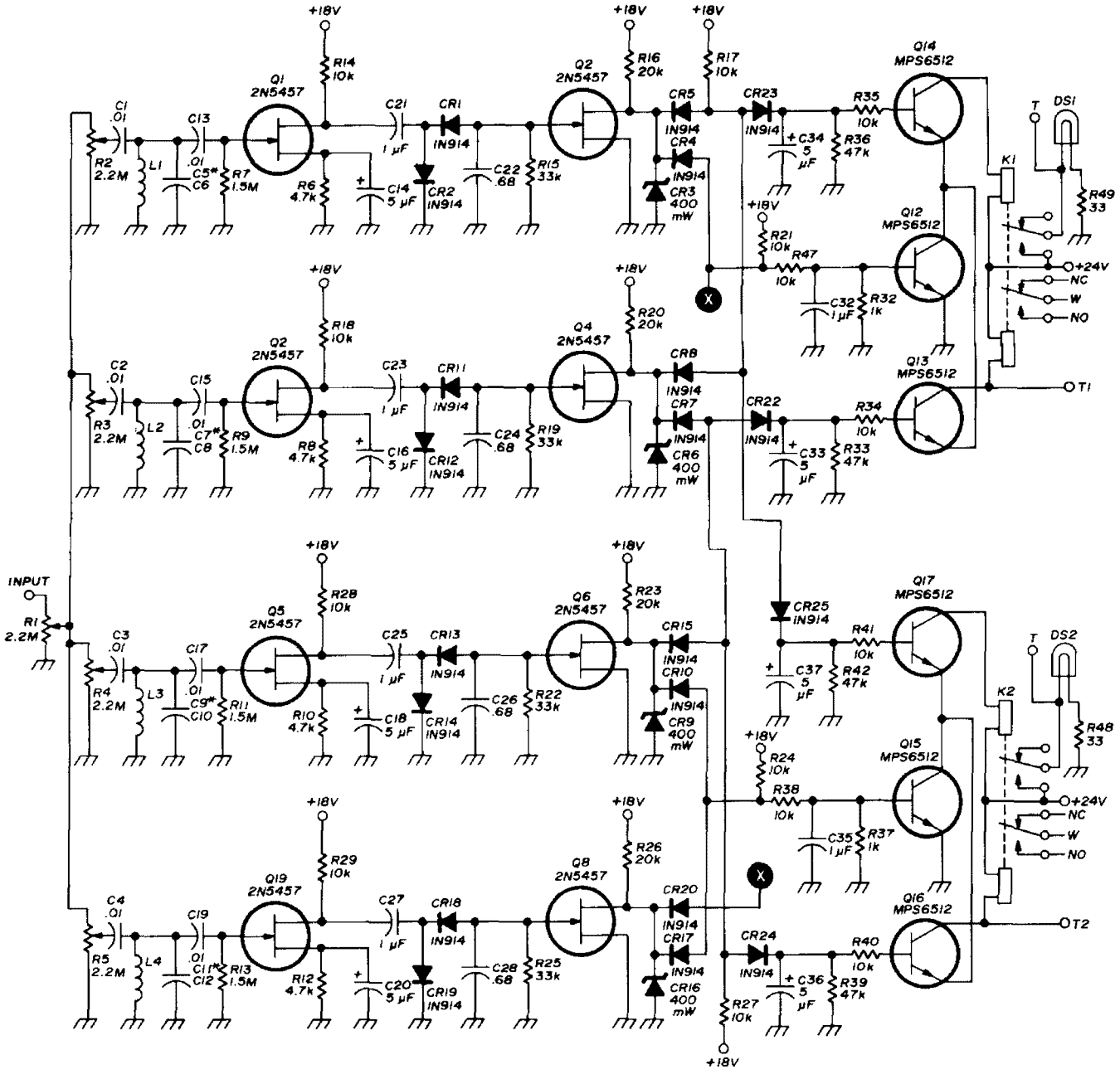


fig. 2. Schematic of the Touch-Tone decoder. Relays K1 and K2 are latching relays such as the Potter & Brumfield FL11D or those available from Circuit Board Specialists.* Capacitors marked with an asterisk are selected for the exact frequency of the desired Touch-Tone tones.

The fet is merely an ac amplifier which boosts the signal up to levels that are easier to work with. The amplified ac signal is rectified by CR1 and CR2, filtered by C22 and is seen on the gate of Q2 as a negative dc voltage. Fet Q2 is connected as a dc amplifier. Therefore, any small change of voltage on the gate will pinch-off the fet which normally conducts the voltage on the drain to ground through dropping resistor R16.

When pinch-off occurs, Q2 ceases conduction and the drain immediately goes positive. This meets one of the requirements of the AND gate made up of CR5 and CR8. Now, go back a little and assume that at the same time tone A was being decoded, tone G was being decoded in the same manner by Q2 and Q4. In this case there would be an immediate rise in

*Etched, drilled, silver-plated printed-circuit boards are available from Circuit Board Specialists, P.O. Box 969, Pueblo, Colorado 81002, $8.50. Relays, latching type, surplus and guaranteed are $2.00.

voltage at the cathode of CR8. This meets the second requirement of the AND gate, allowing the voltage at the junction of R17 and CR23 to go to a positive level. This positive voltage passes through CR23 and is impressed upon the base of Q14 and is stored there by capacitor C34 just long enough to keep transistor Q14 turned on.

Although this all happens when you depress one digit on the Touch-Tone pad, the relay still has not been picked up. To do this you must depress another button and decode two more tones. Let's assume that you depressed the digit 7 (tone A and tone F). Tone A would meet one of the requirements of the gate at diode CR4 while tone F would meet the other requirement at diode CR20, providing a positive voltage at the junction of R21 and R47 which is fed to the base of transistor Q12.

Transistor Q12 will turn on heavily and, in effect, ground the emitter of Q14. This passes all the current in the relay coil to ground, closing the contacts and performing the function. Since relay K1 is a latching relay, it is not necessary to hold it closed. Once latched, the relay stays closed until another signal releases it. This is done in much the same manner as the initial closure. Through different decoding tones a voltage is imposed on the base of Q13 and stored in capacitor C33. This voltage turns on transistor Q12 again, thereby energizing the off coil or relay K1, restoring the relay and the function back to normal.

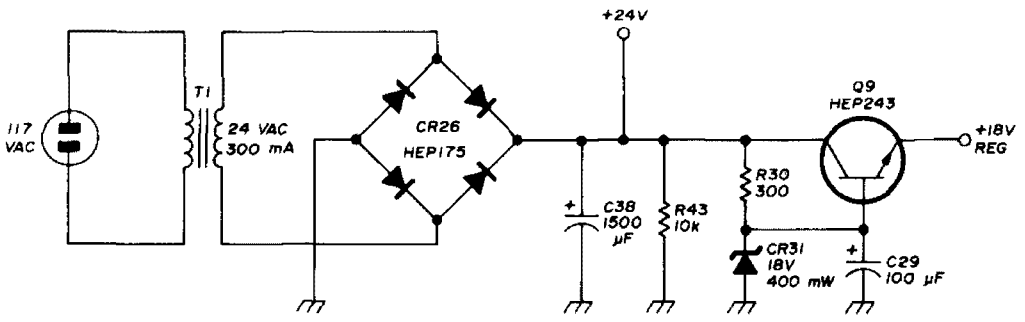


fig. 4. Power supply for the Touch-Tone decoder. Transformer T1 is an Archer 273-1386 (Radio Shack) or equivalent.

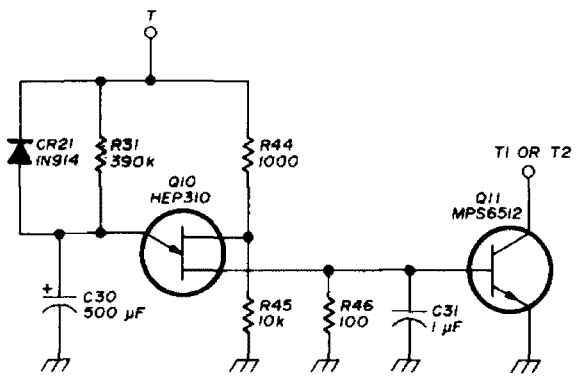


fig. 3. Time-out timer for the Touch-Tone decoder. Resistor R31 is selected for the desired time; value shown is for about 3 minutes.

time-out timer

Suppose that the function which you wanted to perform with your decoder required a time-out timer, such as 3 minutes for an autopatch hookup. This is done very easily and inexpensively by unijunction transistor Q10 in **fig.** 3. When relay K1 closes a voltage on one set of the contacts it turns on PL1, letting you know that it all worked. This voltage is also fed through point T of the relay to point T of Q10. This voltage is dropped through resistor R31 and allowed to gradually build up across capacitor C30. At a certain predetermined level or time lapse transistor Q10 goes into conduction, turning on Q11 for a split second and energizing the off coil of relay K1 thereby timing out the function. Diode CR21 is used for a fast bleed off of C30 so that the timer will be ready for the next time function.

ham radio

two-band antenna matching with stubs

Complete design details for an antenna stub matching system for two harmonically-related amateur bands

Bill Johnson, W6MUR, Box 803, Ben Lomond, California 95005

With the sun spot cycle going down there will be greater interest in 40, 80 and 160 meters for DX work in the next few years, bands where rotary beams are difficult if not impracticable. Wire antennas still have their place, and it is desirable, if possible, to build antennas that will have reasonable directional characteristics on at least two bands. For example, the familiar dipole fed with open-wire line operates as two half-waves in phase on the second harmonic and exhibits slight gain in the broadside direction. The popular W8JK array can also be operated on two harmonically-related bands.

The principal problem with the standard arrangement is that the center impedance of such an antenna at the harmonic becomes quite high, on the order of several thousand ohms, and the bandwidth also narrows; that is, the reactance and resistance change is quite large around resonance. Thus, there is a very high vswr on the open-wire line. While high vswr does not result in appreciable *loss* on a good open-wire line, it does lead to problems in maintaining good balance to ground and in minimizing radiation from the transmission line itself.

High vswr also complicates the antenna tuner in going from one band to another; depending on the length of feeder, it may be necessary to switch

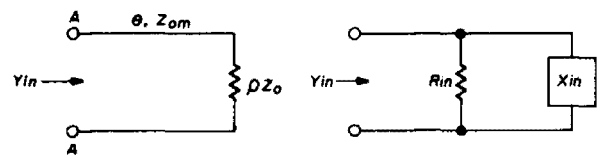
from series tuning on one band to parallel tuning on the other, and band changing becomes complicated. Then, there is the matter of the high voltages and currents along the mismatched line. Even a 600-ohm *flat* line at maximum power will have about 600 to 700 volts rms of rf across it; when the vswr is high this voltage will be appropriately higher. High voltages and currents along a transmission line can cause problems in coupling into other lines (such as telephone lines and tv lead-ins) and in arcing to nearby objects. In short, a low vswr is very desirable, even on an open-wire transmission line.

stub matching

It is possible, using stubs and matching sections, to design an antenna that can be operated on two bands, say 80 and 40 or 160 and 80, so that a reasonable match will be achieved to a specified open-wire line without switching. This particular problem hasn't been treated in the various antenna handbooks where stub matching for only one frequency is discussed.[1]

The analytical approach to this problem can get pretty complicated, and in fact, intractable unless it is done in the right way. The solution lies in using the familiar transmission-line equations in admittance, rather than impedance form, and further, in using half-angle rather than double-angle formulas when going between bands. That is, the electrical length of the matching section and stub is defined at θ at the harmonic frequency, say 7 MHz, and the corresponding length at the fundamental, 3.5 MHz in this case, becomes $\theta/2$.

Fig. 1 gives the basic relationships, the input admittance of a lossless transmission line having characteristic impedance Z_o and length θ. In stub matching, the stub (whether shorted or open) is con-



$$\frac{Z_{om}}{Z_{in}} = Y_{in} = G_{in} + jB_{in} \qquad G_{in} = \frac{Z_{om}}{R_{in}} \qquad B_{in} = \frac{Z_{om}}{X_{in}}$$

$$G_{in} = \rho \frac{1 + (\tan\theta)^2}{\rho^2 + (\tan\theta)^2} \quad (1) \qquad B_{in} = \tan\theta \frac{\rho^2 - 1}{\rho^2 + (\tan\theta)^2} \quad (2)$$

$$\text{at } f \quad G_{in1} = \frac{2\rho_1}{\cos\theta\,(\rho_1^2 - 1) + (\rho_1^2 + 1)} \quad (3) \qquad \text{at } f \quad B_{in1} = \frac{\sin\theta\,(\rho_1^2 - 1)}{\cos\theta\,(\rho_1^2 - 1) + (\rho_1^2 + 1)} \quad (4)$$

$$\text{at } 2f \quad G_{in2} = \frac{\rho_2}{(\cos\theta)^2\,(\rho_2^2 - 1) + 1} \quad (5) \qquad \text{at } 2f \quad B_{in2} = \frac{(\sin\theta)(\cos\theta)(\rho_2^2 - 1)}{(\cos\theta)^2\,(\rho_2^2 - 1) + 1} \quad (6)$$

$$\text{where } \rho_1 = \frac{R1}{Z_{om}} \text{ and } \rho_2 = \frac{R2}{Z_{om}}$$

fig. 1. Basic relationships of open-stub impedance matching at two frequencies, f and 2f.

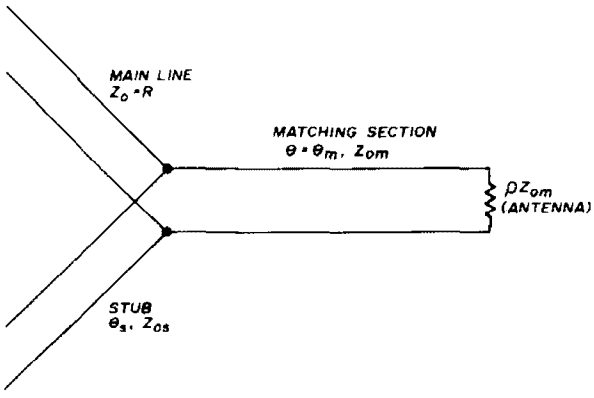


fig. 2. The matching problem.

nected a distance, θ, back from the load or from a known minimum or maximum in the standing wave of the unmatched line. The reactance of the stub is made equal and opposite to the input reactance of the matching section at that point.

For two-band matching the problem is depicted in **fig. 2.** The problem is to find a matching section and stub such that the input impedance with the stub connected, at the point of connection, is resistive and equal to $Z_o = R$ of the main transmission line on each of the two bands. To do this, you must also know the antenna impedance at both frequencies. This is assumed to be resistive, justified on the basis that the antenna will be pruned to resonant length or at least carefully calculated.

Resistance R1 is defined as the antenna resistance at frequency f, and R2 as the antenna resistance at frequency 2f.

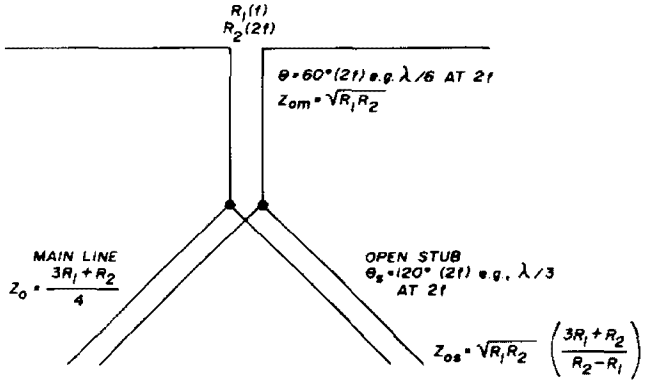


fig. 3. Simplified Johnson match for two-frequency antenna impedance matching.

Using the appropriate trigonometric identities, **eq. 3** through **eq. 6** in **fig. 1** may be derived from **eq. 1** and **eq. 2.** These are the basic relationships for two-band matching. There are four unknowns, Z_{om}, the characteristic impedance of the matching section, Z_{os}, the characteristic impedance of the stub, θ_m, the electrical length of the matching section ($\theta = \theta_m$), and θ_s, the electrical length of the stub.

Resistance R1 has previously been *specified* as the antenna resistance at frequency f. R2 is the antenna resistance at 2f. When the reactance has been tuned out by the stub, the resistance R is equal to

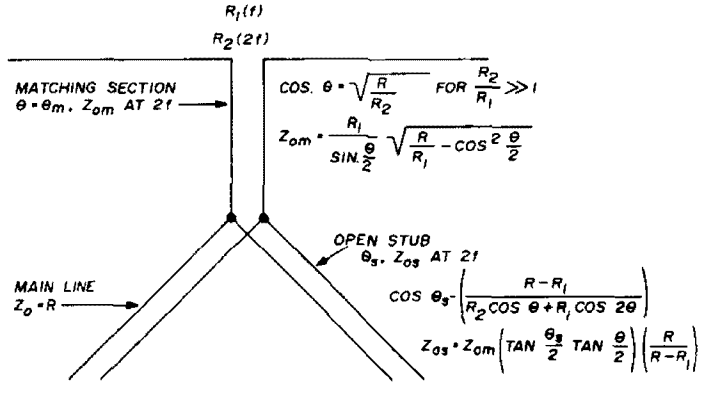


fig. 4. Generalized Johnson match for two bands where R2, the antenna impedance on the higher frequency band, is much greater than R1, the antenna impedance on the lower band, the usual case.

$$R = \frac{Z_{om}}{G_{in1}} = \frac{Z_{om}}{G_{in2}}$$

Where R is the desired match to the main transmission line impedance, Z_o, as shown in **fig. 2.**

special case

Before proceeding to the general case, a most interesting special case is where

$$\rho_1 = \frac{1}{\rho_2} = \rho \text{ or } Z_{om} = \sqrt{R1R2}$$

For this case **eq. 3** is set equal to **eq. 5** to provide the following simple quadratic expression

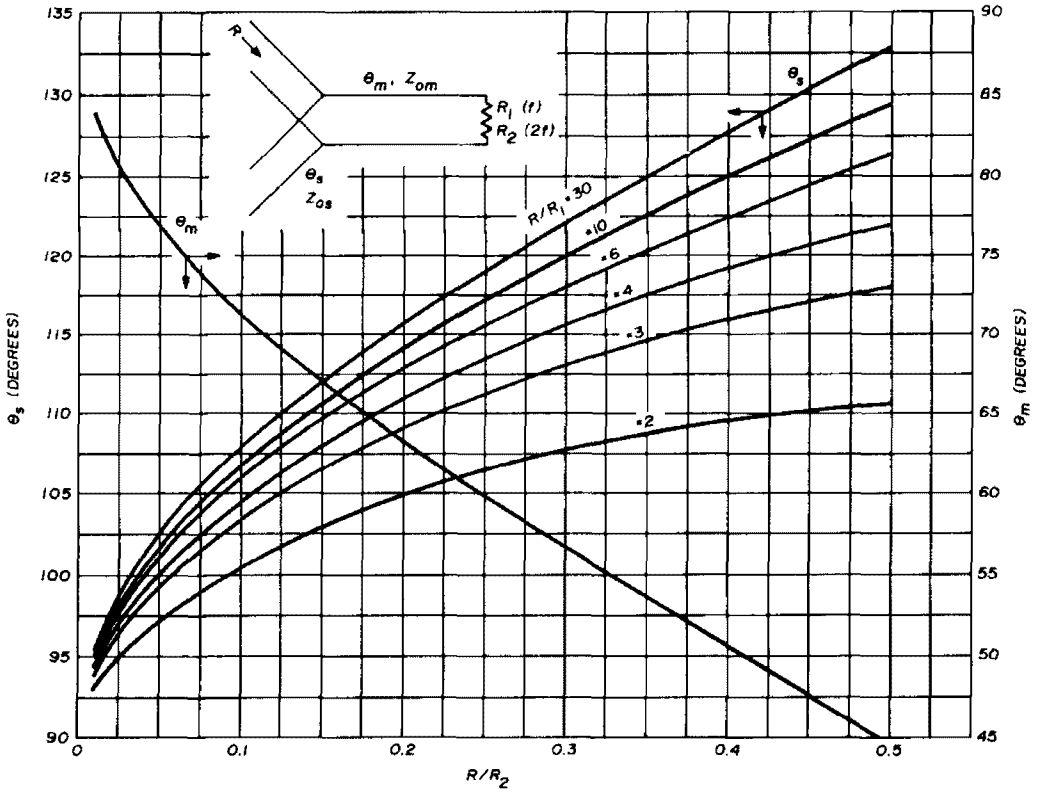


fig. 5. Electrical length of the matching section, θ_m, and the stub, θ_s, at frequency 2f where R2 is much greater than R1.

$$2C^2 + C - 1 = (2C - 1)(C + 1)$$

$$C = \tfrac{1}{2} \text{ or } C = -1$$

where $C = \cos\ \theta$

For this special case $\theta = 60^\circ$ or 180° at 2f. The latter case is the familiar quarter-wave section at f, half-wave section at 2f. The input impedance at points A-A in **fig. 1** is simply R2 at both frequencies. The problem here is that R2 is usually high, so the main line impedance to match it is unreasonably high. The other case, θ =

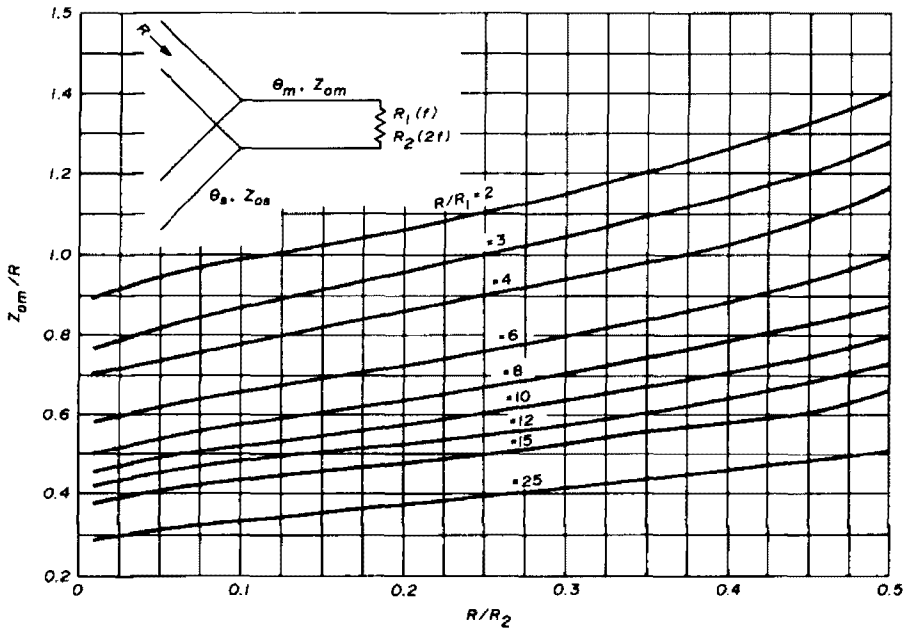


fig. 6. Characteristic impedance of the matching section, Z_{om}, where R2 is much greater than R1.

60° at 2f, is much more useful. For this case the result is

$$R = \frac{3R_1 + R_2}{4} \qquad (7)$$

To tune out the reactance in this $\theta = 60°$ case requires an open stub exactly 120° long at 2f with a characteristic impedance of

$$Z_{os} = 400\left(\frac{3 \times 80 + 2000}{2000 - 80}\right) = 467 \text{ ohms}$$

$\theta_m = 60°$ at 2f (1/6 wavelength)

$\theta_s = 120°$ at 2f (1/3 wavelength)

You would have a perfect match for a 560-ohm line on both frequencies f and 2f by connecting the two 400-ohm and 467-ohm lines as shown in **fig. 3.**

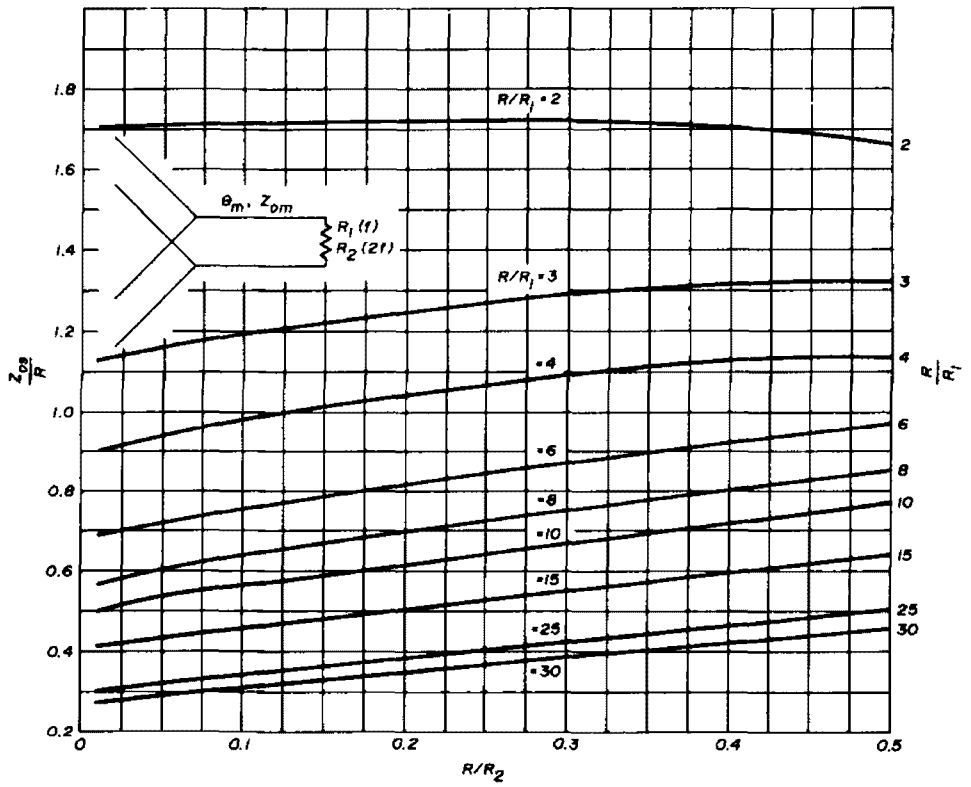


fig. 7. Characteristic impedance of the stub, Z_{os}, where R2 is much greater than R1.

$$Z_{os} = \sqrt{R_1 R_2}\,\frac{3R_1 + R_2}{R_2 - R_1} \qquad (8)$$

The simplified *Johnson Match* is shown in **fig. 3.** Note that the main line characteristic impedance $Z_o = R$ is *not* arbitrary in this case because the quantity Z_{om} has been specified rather than treated as an unknown.

As an example, suppose you had an antenna such that R1 = 80 ohms and R2 = 2000 ohms. Then

$$R = \frac{3 \times 80 + 2000}{4} = 560 \text{ ohms}$$

$$Z_{om} = \sqrt{80 \times 2000} = 400 \text{ ohms}$$

general case

In the special case just considered, you do not have an arbitrary choice of main line characteristic impedance, at least if you want a perfect match. R2 may be considerably higher than the 2000 ohms assumed in the above example, and thus, R may still be unreasonably high, In the general case you first use **eq. 3** and **5** to find Z_{om} and Cos θ (θ and θ_m are used interchangeably here). Then, **eq. 4** and **6** are used together with the stub reactance formula to find θ_s and Z_{os}. The result,

when R2 is much larger than R1, the usual case, is given in fig. 4 as the generalized *Johnson Match.*

To simplify finding the required lengths and characteristic impedances of the line sections, figs. 5, 6 and 7 have been prepared using the approximate formulas given in fig. 4 which are valid when R2 is much larger than R1. In fig. 5 are given the electrical lengths of the

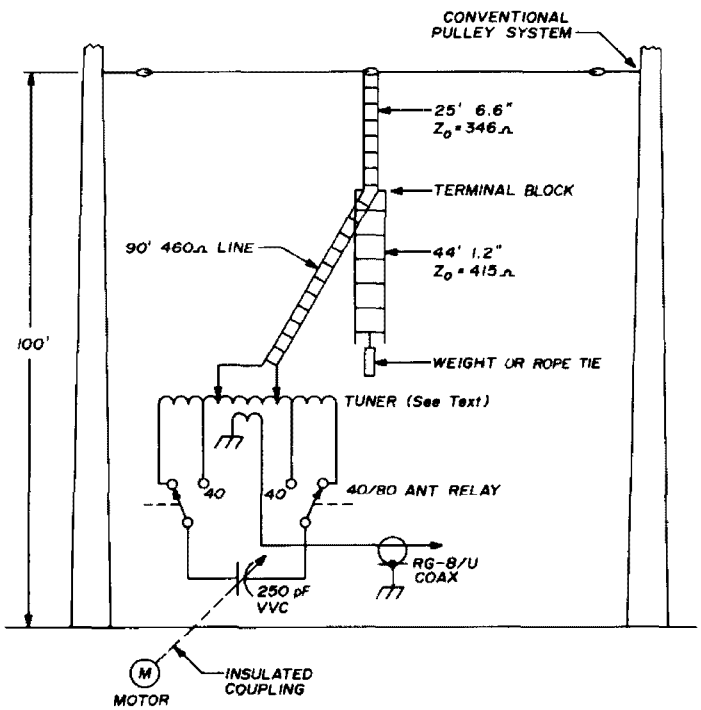


fig. 8. Two-band 3.5- and 7.0-MHz antenna system used at W6MUR incorporates the stub matching system described in the text.

matching section and stub in degrees at the harmonic frequency 2f. In fig. 6 is the Z_{om} family giving the required characteristic impedance of the matching section, and in fig. 7 is the Z_{os} family giving the required characteristic impedance of the open stub.

As an example in using these curves, assume an antenna where R2 = 4000 ohms at 2f and R1 = 60 ohms at f. Also assume you want matching on both frequencies to a 600-ohm line, R = 600. Then, compute R/R2 = 0.15 and R/R1 = 10 and enter the curves to find

from fig. 5, at R/R2 = .15 and R/R1 = 10, $\theta_m = 67.2°$, $\theta_s = 111°$ at 2f;

from fig. 6 find $Z_{om}/R = 0.55$ and compute Z_{om} as 600 x .55 = 330 ohms; and

from fig. 7 find $Z_{os}/R = 0.59$ and compute Z_{os} as 600 x .59 = 355 ohms.

antenna input resistance

An excellent, if somewhat obscure, source for accurate information on the center reactance and resistance of antennas in the vicinity of resonance is given in reference 2. This excellent book, by the way, also has sufficient information in graphical form to enable you to design a very good three-element parasitic beam for either best front-to-back ratio or for maximum forward gain (the two do not coincide) without guesswork as to the lengths of driven element, reflector and director. On pages 20 through 25 of this reference will be found some curves giving the center impedance of "nearly half-wave" and "nearly full-wave" center-fed antennas for various conductor thickness. Additionally, the effects of spaced multiple wires are also given.

The effect of ground must also be considered in estimating antenna resistance. With city-lot installations and low heights, this can get pretty indefinite, but if the antenna is reasonably in the clear, curves such as those given by Kraus[3] can be used. It is noted, for example, that the half-wave antenna, when 0.34 wavelength high, has a resistance not of 73 ohms as in free space, but close to 100 ohms. When the half-wave antenna is low, say 0.1 wavelength high, the resistance drops to something like 23 ohms.

test results

The antenna shown in fig. 8 and 9 has been installed and tested at W6MUR for over two months with excellent results when this article was written. The antenna resistances were estimated for the multiple-wire arrangement at 88 ohms at 3.5 MHz and 2440 ohms at 7.0 MHz, and

the matching section and stub were proportioned accordingly. The antenna tuner is a simple parallel-resonant circuit having an edge-wound ribbon coil (surplus) of 20 turns, 4¼ inches in diameter, 8-inches long (about 18 microhenries), tuned by a 250 pF vacuum-variable capacitor driven through an isulated coupling by a reversible gear motor turning a few rpm. The

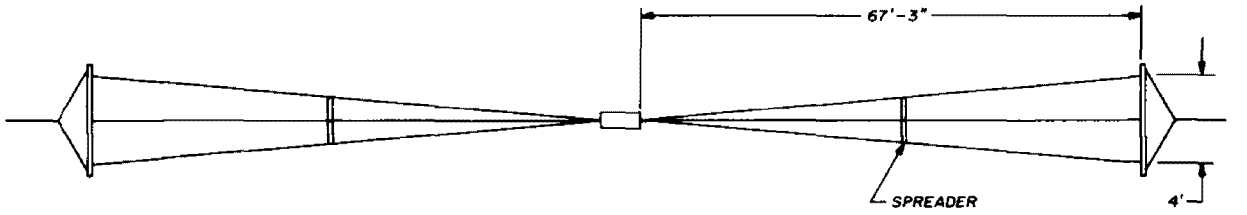
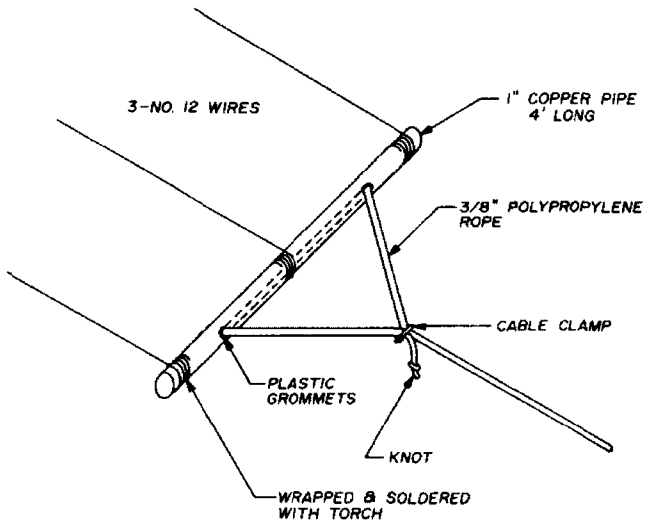


fig. 9. Construction details of the 3.5- and 7.0-MHz antenna used at W6MUR.



antenna was tapped on to the coil and a fixed link constructed of three turns of number-10 wire inside the coil, to which was connected the 50-ohm coax. The same coil was used on both 3.5 and 7 MHz, with the capacitor tapped down on the coil for 7 MHz by means of an antenna switching relay.

Using a grid-dipper signal source and an antenna impedance bridge looking into the link, the points where the antenna feedline was tapped on to the coil were varied until, at resonance, the bridge read 50 ohms at both frequencies. This worked out to be 3 turns between the feeders. The vswr measured into the coaxial line at the transmitter is 1.05:1 at 3500 kHz, 1.6:1 at 3800 kHz, 1.5:1 at 7000 kHz, and 1.9:1 at 7250 kHz. In tuning, once the tuner is very near resonance, you simply watch the reflected power as the motor is energized, and set the variable capacitor for a minimum reading.

This antenna, at 100-feet high, has proved remarkably good for DX work in the preferred direction, which is toward Asia and South America. In the CQ DX contest, November, 1972, on 3.5 MHz 34 countries, 19 zones and 172 stations were worked. The antenna proved itself in the pile-ups for ZD3Z, 9L1GG and 5T5CJ, all worked through eastern-U.S. QRM on 3.5 MHz. Both VU and UL7 have been worked with this antenna on 7-MHz CW, via the long path around sundown.

The main objective of achieving a good match on both bands to the open-wire feedline without excessive switching has been realized. It is hoped that the approach presented here will be useful to others wanting two-band antennas, even if they don't happen to have 135-foot redwood trees to tie them to!

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2. Shintaro Uda and Yasuto Mushiake, *Yagi-Uda Antenna,* Sasaki Printing and Publishing Co., Ltd., 27 Tsutsumi-dori, Sendai, Japan, distributed by Maruzen Co., Ltd., 2-chome, Nithombashi-dori, Chuc-ku, Tokyo, Japan, 1954, pages 20-25.

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ham radio



simple and accurate rf power meter

A useful device for measuring transmitter power output

Adrian Weiss, K8EEG, 213 Forest, Vermillion, South Dakota 57069

The increasing popularity of QRP operation has had one beneficial effect on amateur radio — an increased awareness of the importance of output power versus input power as a measure of performance. It has been generally accepted practice to state the performance of a given circuit in terms of input power capability, but we have long known that input power is no criteria for judging the important aspect of performance; i.e., what the circuit can deliver to the antenna. Our continued use of this archaic approach is perhaps due to the odd FCC practice of defining amateur service power limits in terms of input, while most other services are defined in terms of *effective radiated power.*

efficiency

Whatever the cause, a rather unscientific practice has occurred, namely the rule of thumb assumption that a given circuit will put out roughly 50% of the input power. But such an assumption is in defiance of all known principles of scientific methodology. One simply cannot *assume* that the efficiency of a given circuit is 50%. In fact, the theoretical ideal of a 60% efficiency factor may be

achieved in practice in some circuits, but more likely the efficiency achieved will be considerably below that level. Measurements I have made over the years have revealed that some commercially produced "novice" kit transmitters hardly came out above 25% efficiency, and some were as low as 10%! These results have been corroborated by other experimenters. The lesson here, then, is that input power is no index of what really matters — power delivered to the antenna.

power output meter

Numerous devices for measuring power output have appeared in the literature. The circuit of fig. 1 consists of two parts. First, a dummy load constructed of noninductive, carbon composition resistors is designed for a certain level of power dissipation. Secondly, a sensitive metering circuit samples the peak rf voltage developed across the dummy load through CR1. That voltage is measured through a series dropping resistor on a 200-μA meter. The data is plotted to show μA versus watts. The formula of fig. 1 is used to convert the voltage developed to average power. Accuracy of a high order can be achieved by close attention to accurate measurements of the calibrating voltage, the actual resistance of the dummy load, and the use of a high-quality microammeter.

construction

The photos illustrate the final constructional approach used in this output meter, which is designed for about 60 watts continuous, and perhaps double that for intermittent service. During experimentation, it was discovered that with power levels above a few hundred

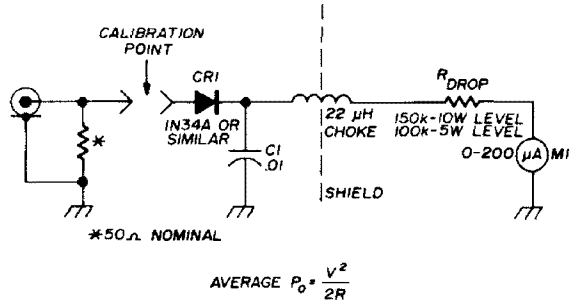


fig. 1. Output power meter schematic. Do not attempt to use a voltmeter at M1. Voltmeters exhibit an impedance which will distort the load seen by the transmitter at the input port at rf, resulting in inaccurate readings due to high swr. However, a high-impedance input meter such as a vtvm may be used instead of the meter shown.

milliwatts, a sufficiently intense rf field was generated, which upset the meter indication accuracy. Hence, a shield was inserted between the meter and the dummy load compartments. Next, it was found that the lead from the rectifying diode provided an rf path to the meter. A small, low-resistance 22 μH rf choke was installed in the lead and positioned halfway through the shield. This eliminated the last traces of rf feedthrough.

If the spirit of this approach is followed — shielding and filtering for rf — an accurate indicating meter will result.

dummy load

The dummy load should be designed for the power levels to be measured. In the meter shown, the dummy load is constructed from 21 two-watt, 1000-ohm resistors in parallel. This load will allow

Author Weiss' opening remarks are well put. Tuning across the ham bands, one often hears remarks such as, "The power here is X watts into an X antenna up X feet." If the power is stated as 1 kW, for example, the fellow on the receiving end might well raise his eyebrows. If the power is stated as dc input (which is often the case), the fellow making the statement is either (a) exceeding the legal power limit for amateur service, (b) his rig is 100% efficient, (c) he's expressing wishful thinking, or (d) he really doesn't understand the difference between input and output power.

True output power is difficult to measure in a radio transmitter, especially if high power is used. This article will permit the measurement of output power with reasonable accuracy for rigs of moderate power. ed.tor

for about 60 watts continuous duty, provided that at least 1/8-inch clearance around each resistor is allowed, and ventilating holes are drilled above and below the load. For power levels under 4 watts, three 180-ohm, two-watt resistors may be paralleled. Whatever setup is devised for the dummy load, an accurate measurement of the resulting actual resistance is necessary to determine the quantity R

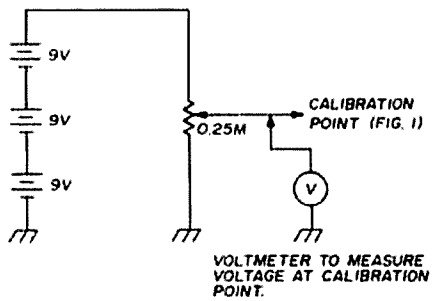


fig. 2. Suggested power supply.

shown in the formula. The dummy load resistors are sandwiched between two sheets of PC board. The lead from one resistor at the center of the board is left about 1/2-inch long, so that it may be inserted directly into the center conductor of the coax receptacle. The ground side of the load is connected to the receptacle outer shield by two no. 18 wires, which pass through both PC boards and go directly to the receptacle.

Even though the dummy load shown in the photo is approximately 3 x 2-5/8 inches, reactance is negligible. When fed through an odd length piece of coax, the swr is not above 1.05:1.

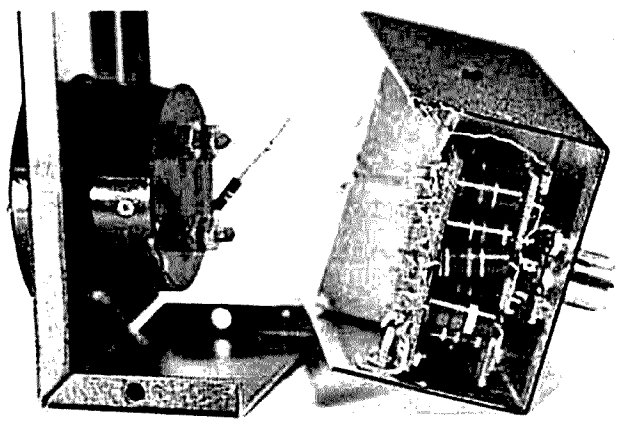
calibration

As noted above, meter accuracy depends upon care in calibration and using a high-quality meter. The best method of calibration is at dc using a variable-voltage power supply. If no such power supply is available, a suitable substitute is shown in fig. 2, and consists of four 9-volt transistor batteries in series to provide approximately 37.5 volts. This supply will calibrate up to about 14 watts. In calibrating, the voltage drop across CR1, the rf choke, and the dropping resistor is automatically accounted for by connecting the calibrating voltage source to the dummy-load side of CR1 – disconnect CR1 from the load first, of course!

Our objective in calibration is to plot the reading in microamps against the voltage at the calibration point (fig. 1). The series dropping resistor value can be determined by experimentation. With the meter used here, a 100k resistor provided just over full-scale meter deflection at 10 watts; a 150k resistor at 5 watts. Table 1 lists the calibration points in the meter shown, and the formula used to calculate the voltage versus watts at chosen power levels. Once the complete number of power levels has been calibrated against microamp readings, the accuracy can be cross checked by attempting to reset the calibrating voltage to given levels by using the newly calibrated microammeter. Results will give you a practical idea of what to expect in the way of accuracy. With care, at least ±5% should be easily achieved. Remember we are ultimately dealing in terms of signal levels, where 1 dB is the lowest possible noticeable difference in signal strength. Accuracy of ±15% is acceptable in these terms.

other applications

The instrument is quite simple and is a valuable addition to any station. Its uses go beyond simple power output measure-



Inside view. Dummy load, coax receptacle and shield. CR1 is to the right of the dummy load PC board. C1 is partially obscured at the rear of the load. Meter and series dropping resistor are at left.

ment. It can be used to calibrate the popular Breune type in-line power output meter, such as was described by DeMaw.[1] Also, it can be used to measure the output of an exciter to determine exactly how much attenuation is needed to provide proper excitation to a linear. For the QRPP operator, such a device is absolutely essential. For the QRO gang, it can provide a much needed moment of truth

table 1. Calibration points calculated from formula of fig. 1 transposed to solve for V^2, i.e.: $V = \sqrt{P_o \times 2R}$ (102 ohms)

P_o (watts)	V (volts)	μA (my meter)
10	31.93	203
9	30.02	194
8	28.50	185
7	26.74	169
6	24.7	158
5	22.5	145
4	20.17	129
3	17.56	113
2	14.28	93
1	10.99	67
0.9	9.58	63
0.8	9.03	59
0.7	8.44	55
0.6	7.82	51
0.5	7.14	48
0.4	6.38	43
0.3	5.53	38
0.2	4.51	30
0.1	3.19	22
0.075	2.76	17
0.05	2.25	14
0.025	1.59	10
0.01	1.01	5

when connected to the pride and joy that's supposed to be putting out (assuming 50% efficiency, that is) 90 watts. The device requires such little time and effort to construct and calibrate, there is little reason why any station should be without it. It provides an indication of the only significant performance factor in the transmitting system — the amount of power the transmitter can deliver to the antenna.

reference

1. Doug MeMaw, "In-Line RF Power Metering," *QST*, December, 1969, page 11; "The QRP 80-40 CW Transmitter," *QST*, June, 1969, page 16.

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circuit improvements for the advanced frequency scaler

Several advanced circuit improvements for the vhf divide-by-ten frequency scaler previously described in *ham radio*

F. Everett Emerson, W6PBC, Belmont Spectrum Research, Belmont, California

This is a follow-up to the divide-by-ten frequency scaler described in the September, 1972, issue of *ham radio.*[1] That article was originally prepared in late 1971 when the Fairchild 95H90 IC was relatively new, and experience with it, ham-wise, was quite new indeed. More than a year's experience with the 95H90 is now behind us, so a later look seems appropriate.

To the serious experimenter, the original article presented no real problems and many letters I received indicated that the device was a worthwhile project and performed as described. Others, being perhaps less knowledgeable, have had some difficulties. Drawing upon the experience of *Belmont Spectrum Research* in its commercial manufacture of scalers, and its experience wtih a great many ICs, as well as letters from users, indicates that attention to the following points should greatly assist in smoothing out difficulties and in making your scaler a truly useful device.

power supply

Experience has shown that *most* 95H90s have optimized at between 4.75 and 4.85 volts. This means that the power supply should provide this voltage to the IC. Accordingly, and this is done in the

commercial version, a different power supply is now recommended for new construction (see fig. 1). This circuit is much more simple than the original and can easily be optimized. The circuit used Fairchild's 7805 voltage regulator IC (National's LM309 is equally suitable) which is rated at 5.0 volts output. These regulators may vary slightly from the "typical" 5.0-volt rating specified on the data sheets, so measure their output voltage. When using them, shoot for 4.85

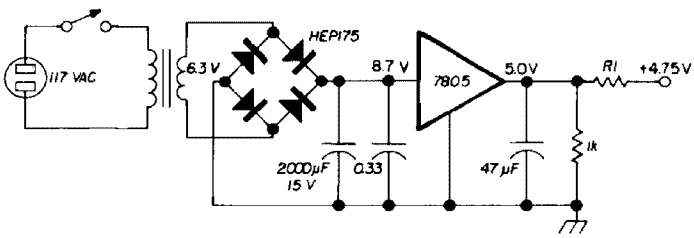


fig. 1. Recommended power supply for the 95H90 frequency-scaler IC. The supply voltage is connected to pins 4 and 5 of the 95H90. If the voltage is greater than 4.75 to 4.85 volts, install resistor R1 (2.7 to 3.9 ohms) and adjust value, as required.

volts at pins 4 and 5 of the 95H90 by inserting, if necessary, a 2.7- to 3.9-ohm resistor in series (R1 in fig. 1).

frequency limit

Experience with many manufactured scalers has shown that all 95H90s are far from being identical as far as their *upper* frequency limits are concerned. Some will go to 325 MHz, many will go to 300 MHz, but some will not go above 260 MHz. They are specified to 220 MHz as a minimum and to 270 MHz as "typical" by Fairchild.

The reason for the wide upper frequency range found in practice is that this IC is heat sensitive. That is, a large temperature rise lowers the upper frequency limit. Hence, the 95H90 should have a good heat sink if its maximum possibilities are to be exploited. The grounding of all unused pins (except pin 14, more on this later) will help. A good commercially made heat sink is made by IERC (their part numbers DC000080B and LIC 214A2WCB). You can make your own heat sink, however, by using a two-inch square piece of aluminum in contact with the top of the IC and bolted to the circuit board.

circuit-board design

A good ground plane is essential in the frequency scaler. This is why, in the original article, lines were removed from a "ground plane" copper board rather than using interconnecting traces. Don't forget you are dealing with very high frequencies where miscellaneous circuit paths may lead to feedback or ground loops and cause instability. Those builders who have used "traces" have had real problems and have ended up remaking their circuit boards.

preamp

If a preamp is used, and one is very worthwhile (the commercial version uses two stages), the 95H90 must be adequately decoupled from the preamp. This is rather simply accomplished by the 22-ohm resistor between the 5-volt supply and the junction of the 180-ohm resistor and the peaking coil shown in fig. 5 of the original article. Do not omit this precaution against feedback or you will end up with an oscillator, not a preamplifier.

low-impedance input

Experience has shown that *most* 95H90s respond best to a low-impedance input. This was not so with the prototype ICs I used for the original article. Thus, it is now recommended that a 68- and

200-ohm bias divider be used for the 95H90. Also, if you would like a further refinement (the original was somewhat of a compromise in order to make it work for the vast majority of cases) the use of a *threshold* control, as shown in **fig. 2**, may be more acceptable for your applications. Try it, you may like it.

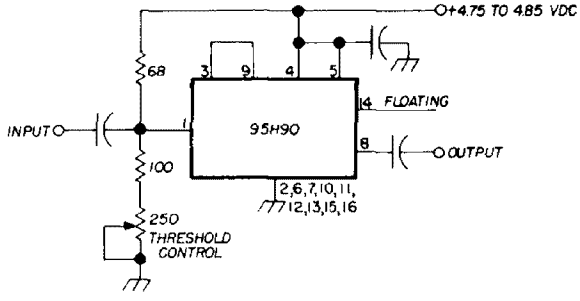


fig. 2. Improved low-impedance input circuit for the 95H90 IC. The 250-ohm threshold control is optional but should prove useful for most applications (see text).

bypassing

Bypassing pins 4 and 5 of the 95H90 (assuming you followed **fig. 3B** of the original article) is very important. The by-pass capacitor should be connected as close as possible to pins 4 and 5. In other words, *zero* lead length!

ground unused pins

Experience has shown that *all* unused pins of the 95H90 with the exception of pin 14, should be grounded. Thus, if you use a negative grounded board as in **fig. 3B** of the original article and use pin 1 as the input, pins 2 and 16 should be grounded. Pin 14 should be left floating (ungrounded). The first few ICs I tried did not require this. However, subsequent experience has proved the efficacy of the grounding of pins 2 and 16. Ground them!

heat dissipation

The 95H90, which dissipates approximately one-half watt continuously, should feel only slightly warm to the touch if operating properly. If it runs hot you have something wrong. Check the voltage at pins 4 and 5. It should not be over 5.0 volts and preferably, only 4.75 to 4.85 volts. Also, check your output resistance (with the unit turned *off*). It should not be much less than 800 ohms at pin 8. If it is less, you undoubtedly have a circuit error.

base diagrams

Be sure you have not somehow confused top and bottom views of the 95H90 and the 2N5179. A number of builders have interchanged them, much to their sorrow. *Always* check a data sheet which shows the basing diagram. In fact, double check it.

TTL interface

Unless you are sure that you need a TTL interface (as shown in fig. **6** of the original article) omit it. There are presently no known counters which require it. If you do use it, however, pay no attention to the V_{cc} and V_{ee} markings on fig. **6**. The proper voltage is already taken care of in the 95H90.

cable termination

In using a scaler with a highly sensitive high-impedance frequency counter, false counting can be experienced if the interconnecting cable between the scaler and the counter is not properly terminated at the counter end of the line. A line termination consisting of a carbon resistor, equal to the characteristic impedance of the line, should be connected across the line at the counter end. This precaution is often overlooked by hams but is standard practice in industry. Tektronix makes a fine termination adaptor for use with 50-ohm coax (their part no. 001-0049-01).

reference

1. F. Everett Emerson, W6PBC, "Advanced Divide-by-Ten Frequency Scaler," *ham radio*, September, 1972, page 41.

ham radio

improved half-wave rectifier

Circuit details of an improved half-wave, choke-input power supply circuit that offers several advantages over conventional circuits

John T. Bailey, 86 Great Hills Road, Short Hills, New Jersey 07078

Suppose you needed a simple power supply in a hurry that would give you 50 volts dc at, say, 100 mA, well filtered, with load regulation at least equal to that of a choke-input filter, and without excessive peak rectifier currents. You might say, "That's not difficult." A conventional full-wave center-tap circuit feeding a choke-input filter would meet these requirements and the parts should be available in the junk box.

You start looking for a transformer with a 120-volt center-tapped secondary in order to get 50 volts dc at the input to the filter choke. Or, if you used a bridge rectifier, you would need a 60-volt untapped secondary. My guess is that you won't find either transformer in your junk box.

half-wave circuit

That being a dead end, let's make it more difficult. How would you design for the same requirements without using any transformer? I'll give you a hint. Try a half-wave rectifier feeding a choke-input filter. But, you say, "This is never done. The regulation would be horrible and the output voltage would be practically unpredictable."

One simple diode added in the right place can fix all that. Simply wire a diode, reversed polarity connection, from the choke input to the common line. That's all. The load regulation will be

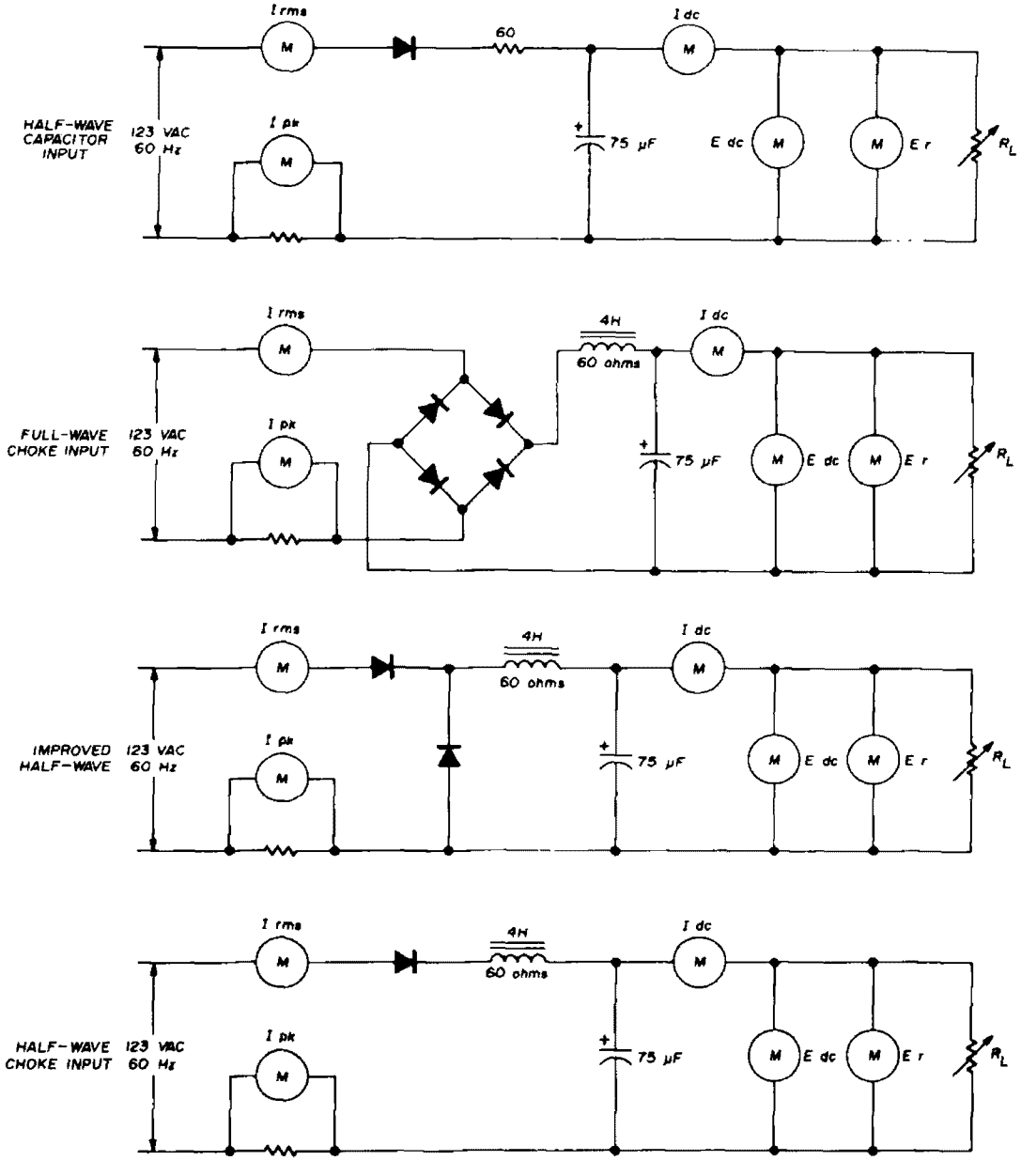


fig. 1. Power-supply circuits measured to show performance of improved half-wave circuit relative to other popular rectifier circuits.

equal to that of a full-wave, choke-input circuit. The output voltage will be one-half that of a full-wave, choke-input circuit or 45% of the rms ac input.

What about disadvantages? There must be some. A minor one is that when the dc output current must be varied from full load to a much lower output current the minimum load current must be greater than that for a full-wave choke-input circuit to prevent the circuit from reaching criticality.

Ripple in the output voltage will be the same frequency as the supply. Its magnitude will be more than for the full-wave, choke-input circuit but much less than for the conventional half-wave, capacitor-input circuit. The ripple voltage waveform will resemble the "full sine wave" ripple of the full-wave, choke-

input circuit rather than the "triangular" waveform of the conventional half-wave circuit.

tests

To delineate the performance of the improved half-wave circuit, measurements were made on the four circuits shown in fig. 1. The results are plotted in fig. 2. Table 1 lists other data applicable to the four circuits using the component values shown for each. If some of these data appear to differ from those derived from theory remember that the data here are measured values on circuits whose components are not the ideal components frequently assumed in textbook analyses.

In these circuits the I_{rms} line currents were measured with a thermocouple milliammeter. The peak line currents were measured with an oscilloscope across a 1-ohm resistor in the line. The ripple waveforms were observed on a scope across the load resistor and the rms values were calculated.

Note particularly that the slopes of the curves for the improved half-wave circuit and the full-wave, choke-input circuit are straight and parallel over their useful spans indicating the internal resistance of each is constant and equal to the other.

table 1. Data applicable to the four rectifier circuits shown in fig. 1.

	half-wave capacitor input	full-wave choke-input	improved half-wave	half-wave choke-input
Ripple voltage, rms				
at full load	9.6	.33	1.2	1.9
at 50% full load	5.5	.31	1.1	1.2
at 20% full load	2.7	.30	1.0	.7
Internal resistance, ohms				
at full load	250	66	66	1,100
at 50% full load	300	66	66	1,700
at 20% full load	500	66	66	3,400
I_{rms}/I_{dc}				
at full load	2.0	1.00	.8	1.3
at 50% full load	2.3	1.05	.8	1.5
at 20% full load	2.6	1.10	1.0	2.0
I_{pk}/I_{dc}				
at full load	5.0	1.10	1.3	2.2
at 50% full load	6.3	1.23	1.5	2.8
at 20% full load	9.2	1.67	2.3	3.8

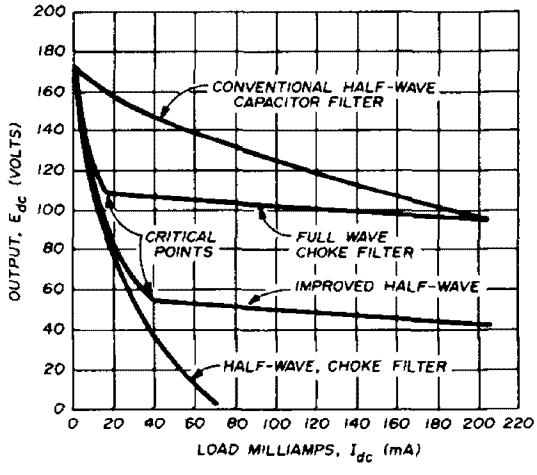
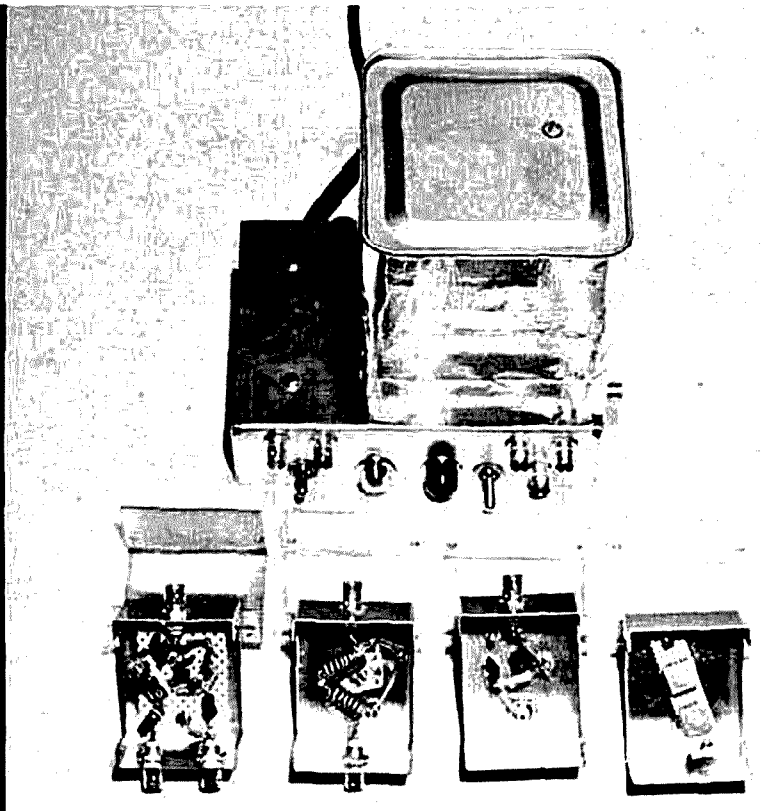


fig. 2. Voltage regulation curves of the rectifier circuits shown in fig. 1. Note the superior regulation of the improved half-wave circuit relative to the other half-wave circuits.

summary

Of course, this improved half-wave circuit is not limited to line operation without a transformer. It is a handy circuit to keep in mind whenever an available secondary transformer voltage is about twice as high as you need to get a certain dc voltage when using conventional circuitry. For instance, a 24-volt filament transformer feeding this circuit will give you a handy 10 volts for those transistor projects where better regulation isn't justified.

ham radio



photos by WA3MEJ

accurate frequency measurement of received signals

How to use an offset frequency measurement system that does not impair the performance of your receiver

Jeff Walker, W4AAD, Pasadena, Maryland 21122

Several articles dealing with frequency measurement of received signals have appeared in the amateur publications.[1-6] The techniques described have been essentially of four types, manual analog comparisons, receiver dial readouts, received signal synthesis and multiplexed up/down counting. Each of these techniques has advantages and disadvantages.

Manual analog comparisons (zero beating and interpolation) are easy and inexpensive to implement but require considerable operator skill and are subject to human error at best. Receiver dial readouts are fine except that they read the frequency to which the receiver is tuned and not that of the received signal. Also, most dial readout systems described to date have a resolution of only 100 Hertz.

Received signal synthesis is good except that the synthesized signal tends to re-enter the receiver front end and cause oscillation problems. The multiplexed

up/down counter approach works well but requires some moderately complex digital design and pretty well eliminates the use of a general-purpose frequency counter.

My goals when designing this measurement system were to obtain precise readout of the received signal carrier frequency (within counter resolution) with complete independence from receiver tuning. The system would offer easy calibration and require no operator-performed analog comparisons, such as zero beating. It would provide stable, non-oscillatory operation and use an existing general-purpose frequency counter. Also, it would have high sensitivity and be capable of operating on all the hf and vhf amateur bands within the limitations of the counter.

To meet all the requirements, I decided to use a modified form of received signal synthesis. Henceforth, I will refer to this technique as *offset counting.* Signal synthesis satisfies most of the design goals, however, it does not satisfy the requirement for stable, non-oscillatory operation. This problem is worsened when a general-purpose counter is used because the synthesized signal must be fed into the counter with coaxial cable, thus allowing some signal leakage. Furthermore, most counters generate at least 1 volt, peak-to-peak, of counted signal and radiate some portion of that. The net result is anything but a stable, non-oscillatory system. The only way to make such a system work is to desensitize it. This compromises the goal of high sensitivity.

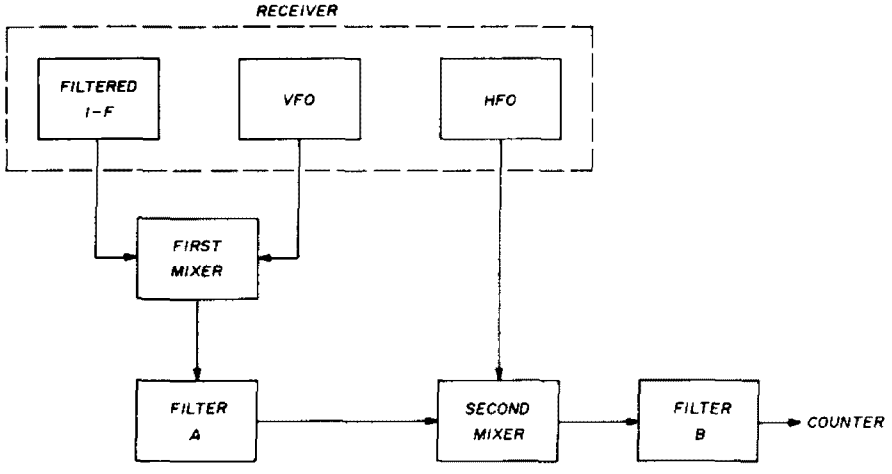


fig. 1. Block diagram of a system to synthesize a received signal for frequency counting.

However, if the counted signal is on some frequency other than the received frequency, high sensitivity is not compro-

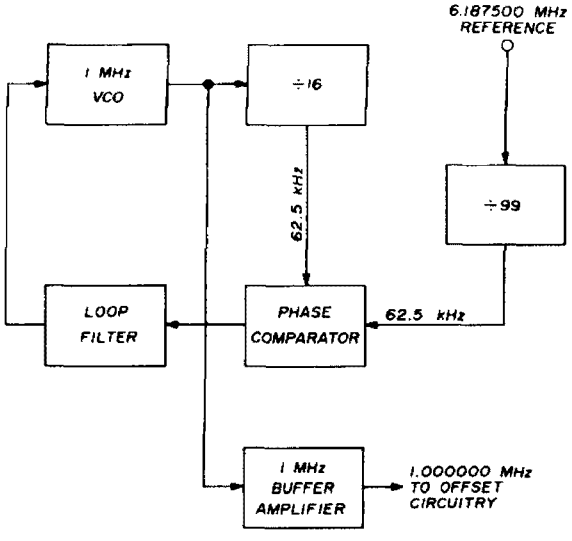


fig. 2. Block diagram of a circuit to phase lock an offset oscillator to an "odd-ball" reference frequency.

mised, and the system provides stable, non-oscillatory operation. The problem here is that you must know *exactly* how the counted signal relates to the received frequency. I accomplish this in the offset counting system by introducing an offset which is derived from the time base used to control the counter gate.

This system was initially designed and used for the high-frequency bands. With the addition of just one mixer and a filter, the technique is expandable to vhf.

theory of operation

To generate a synthesized received signal for counting, the system shown in fig. 1 would be used. Such a system, typically, would be prone to oscillation. I might point out that this system can be made to work by enclosing the mixers, filters and counter section in the same small, well shielded and filtered enclosure. For this to work, after the signal has been synthesized, absolutely no rf can be allowed to escape from the enclosure.

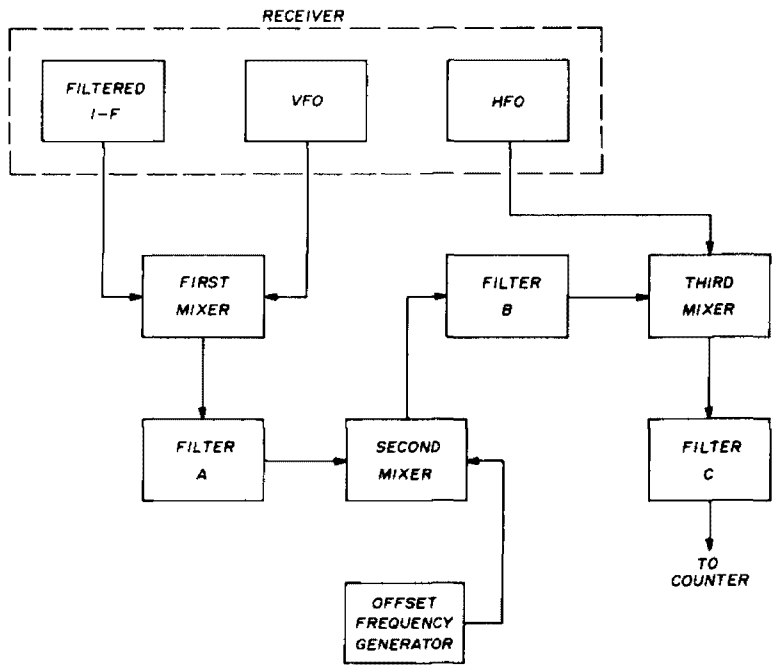


fig. 3. Simplified block diagram showing how an offset counting system may be used to measure the frequency of a received signal.

This is not an easy task to accomplish; shielding and filtering well in excess of usual amateur practice would be required to build a truly stable, high sensitivity system. In fig. 1, the first mixer combines the filtered i-f output with the variable-frequency oscillator to produce sum, difference and spurious products.

Depending on the mixing scheme of the receiver, filter A is designed to pass only the desired product and attenuate all others. The output of filter A is then fed to the second mixer where it is combined with the output of the high-frequency oscillator. Either the sum or difference component is selected by filter B (depending on the receiver mixing scheme) and fed to the counter. This diagram is somewhat simplified in that buffer amplifiers may be required between the receiver oscillators and external mixers as well as between the first and second external mixers.

For example, when designing a synthesis system around a Collins 75S-3 receiver, the hfo operates 3.155 MHz above the low end of the band in use, the first i-f is a passband from 2.955 to 3.155 MHz, the vfo tunes from 2.7 to 2.5 MHz, and the second i-f is 455 kHz. The first mixer uses difference mixing and the second mixer uses summation mixing.

Thus, in fig. 1, the first mixer sums 455 kHz and 2.7-2.5 MHz. The desired output of this mixer is 3.155-2.955 MHz (3.155 MHz is at the low end of the band being tuned). This range is selected by filter A which may be either a bandpass or highpass type.

The second mixer takes the difference between 3.155-2.955 MHz and (F_1 + 3.155 MHz) (F_1 is the frequency of the

low end of the band being used, for the 7.0-7.2-MHz band [F_1 + 3.155 MHz] is 10.155 MHz). The difference output of the second mixer is then exactly the same frequency as the received signal. Filter B selects this product.

To modify the synthesis system to offset counting, an offset frequency must be inserted which will displace the counted frequency sufficiently from the received frequency that the receiver will not respond to it. This offset can be any value large enough to get the counted frequency out of the i-f passband, or preferably, out of the i-f *and* rf passbands. The two absolute conditions which must be placed on the offset frequency are that it be stable (preferably locked to the counter time base) and that it be known to an accuracy at least as good as the highest resolution expected from the counter.

For example, if some multiple of the counter time base is 6.1875 MHz, this would be a perfectly acceptable offset frequency. However, it would be inconvenient to have to add or subtract this number from every frequency you measure. A much more convenient number would be 1.000 or 5.000 MHz. If a nice, round offset frequency is not directly available, it may be possible to phase lock a lower stability signal.

In the case of the 6.1875-MHz signal mentioned above, it is possible to use the circuit block diagrammed in **fig. 2** to phase lock a 1-MHz oscillator to the 6.1875-MHz signal. The phase locked 1.000-MHz signal may then be used for the offset frequency. The reference frequency may be any value that has a common denominator with the offset frequency with which you desire to phase lock.[7]

To see how the offset frequency would be added to the basic synthesis system, refer to **fig. 3**. The offset counting system requires one additional mixer, another filter and an offset frequency source. The signal flow is the same as in **fig. 1** up to the second mixer. At this point the offset frequency is combined with i-f/vfo sum. Either the sum or difference product of the second mixer may be selected by filter B.

In general, the sum product would be preferred as this will ease the design requirements on filter C (in the case of the 75S-3 mixing scheme, it also allows 10-MHz counters to operate up through 20 and 15 meters). The output of filter B is combined with the receiver hfo in the third mixer. Either the sum or difference

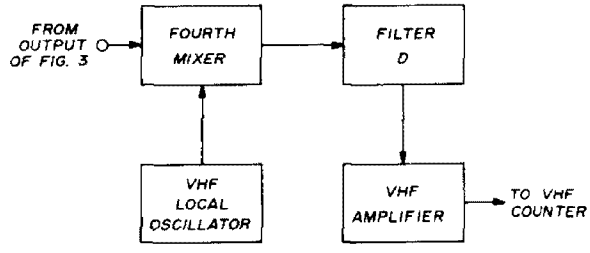


fig. 4. Offset frequency measuring system may be extended to the vhf bands with the setup shown here.

may be selected by filter C; the difference product would be preferred as it will result in a lower frequency to be counted and, thus, be within the range of more general-purpose counters.

Note that I did not combine the receiver i-f, vfo and hfo in sequence; to have done so would have resulted in the generation of the undesired received signal. Instead, the offset was entered before the hfo signal.

The offset counting system may be extended to include the vhf bands by adding an additional mixer, filter and source of vhf local-oscillator signal. Refer to the block diagram in **fig. 4** to see how this is accomplished. The output from filter C in **fig. 3** is fed to the fourth mixer along with a portion of the vhf converter local oscillator signal. Depending on whether the local oscillator is above or below the received vhf signal, filter D should be designed to select the difference or sum output, respectively, of the fourth mixer.

It is likely that the output from filter D will be inadequate to drive your vhf

counter so a stage of amplification will probably be required to raise the signal level. Don't forget that the offset frequency is still in the system and the vhf counter reading will include this factor.

At this point the advantage of using 5.000 MHz or 1.000 MHz for the offset frequency should be realized. Not only will the counter correction be easy to manipulate mentally, filtering out the image from the fourth mixer will be much easier if it is 10 or 20 MHz from the desired signal, rather than some considerably smaller value.

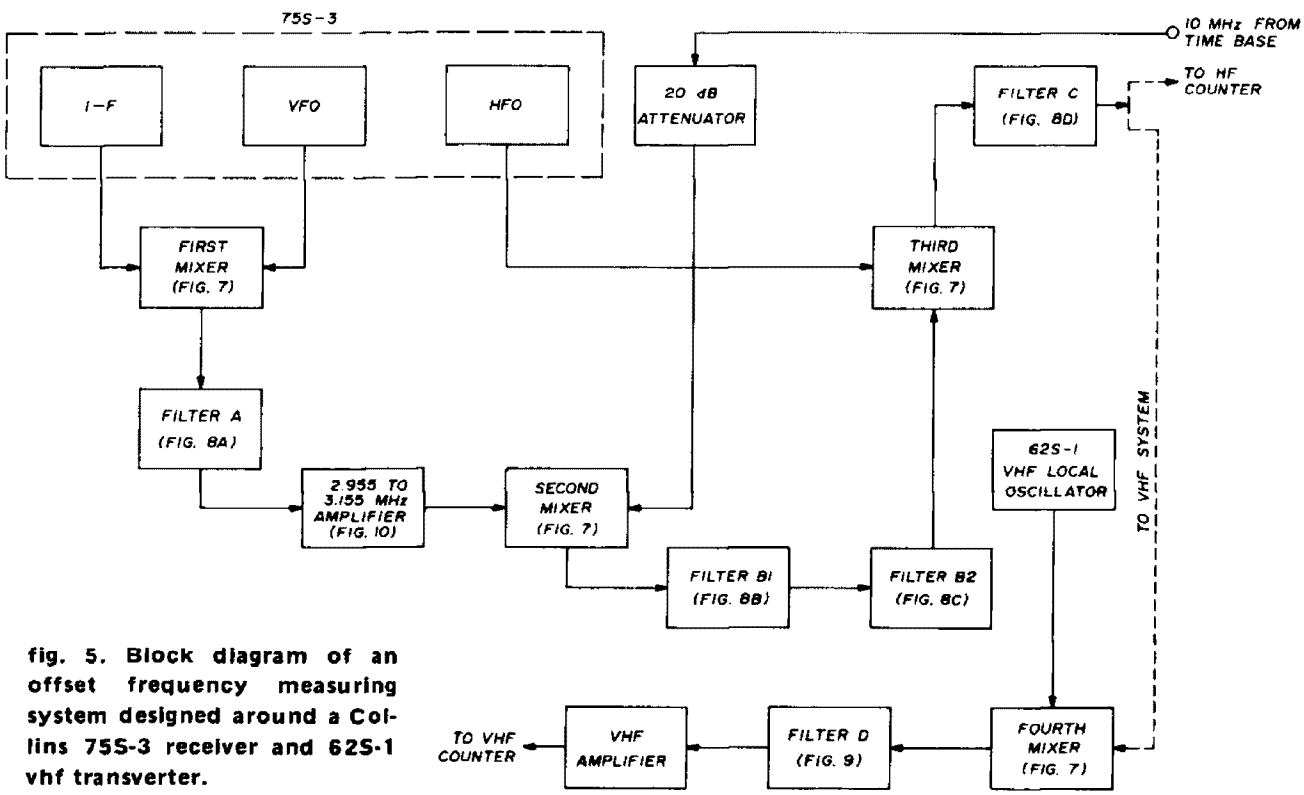


fig. 5. Block diagram of an offset frequency measuring system designed around a Collins 75S-3 receiver and 62S-1 vhf transverter.

construction

As can be seen in the photographs, I used a modular format in building this system. Each block of figs. 3 and 4 is a separate module with the exception of the offset frequency source which is a part of my counter time base. This modular approach allowed me to mount the entire frequency measuring system in a convenient, long, narrow space on my operating table, behind the transmitter and receiver.

This is desirable since the cables feeding the oscillator signals from the receiver and vhf converter can be kept to a minimum length. However, there is no reason why the system can't be built into one large enclosure. If your operating table will permit the inclusion of an additional moderately sized unit you may wish to follow this approach. Just be sure to provide good shielding between the various mixers; otherwise, undesired signal leakage around the mixers will occur.

The modular approach also allows the builder to tackle construction of one distinct portion of the system at a time. Each module can be built, tested and set aside for later inclusion in the system. A good compromise between the modular and single unit approaches would be to build the separate modules and then assemble them in a mainframe or chassis; that way adequate inter-module shielding will be assured.

system description

My frequency measuring system was designed around a Collins 75S-3 receiver,

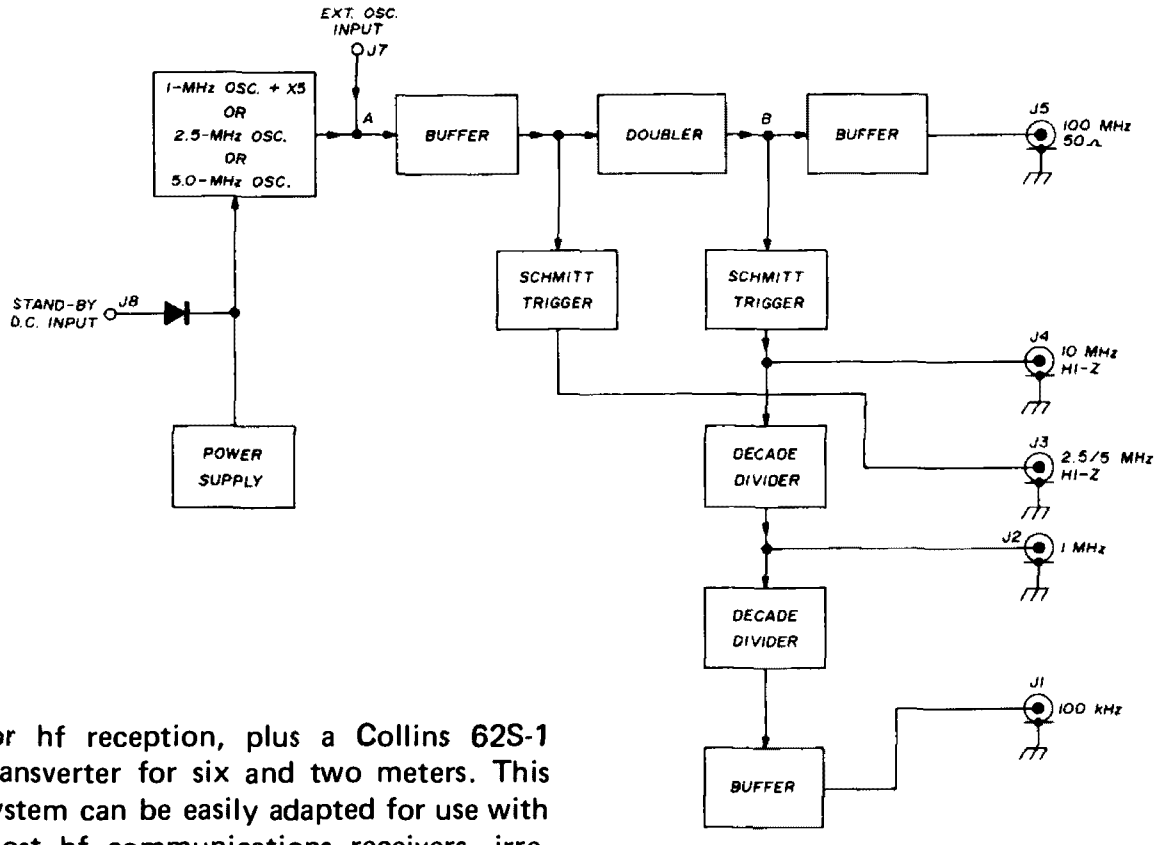


fig. 6. Block diagram of the frequency-counter time base and offset frequency source.

for hf reception, plus a Collins 62S-1 transverter for six and two meters. This system can be easily adapted for use with most hf communications receivers, irrespective of mixing schemes or number of conversions. Single-conversion receivers will require one less mixer, and triple conversion units, one more.

Assuming the modular construction approach is to be used, it is obvious that there is no provision for bandswitching, at least in the classic sense. To change bands requires changing one or more of the filter modules. This is no handicap unless you are an ardent band-hopper or want to measure the exact frequency of every station you work during a DX contest! Even then, you are in luck, providing you hop around the right bands; more on this later.

Complete block diagrams of the frequency measuring system implemented around the 75S-3/62S-1 combination are shown in figs. 5 and 6. In fig. 5, the 75S-3 i-f and vfo signals are combined in the first mixer. The desired output from this mixer is 2.955-3.155 MHz. This is selected by filter A which is a nine-section highpass filter with a cutoff frequency of 2.9 MHz. The output of this filter is fed to a tuned rf amplifier providing approximately 25-dB gain over the

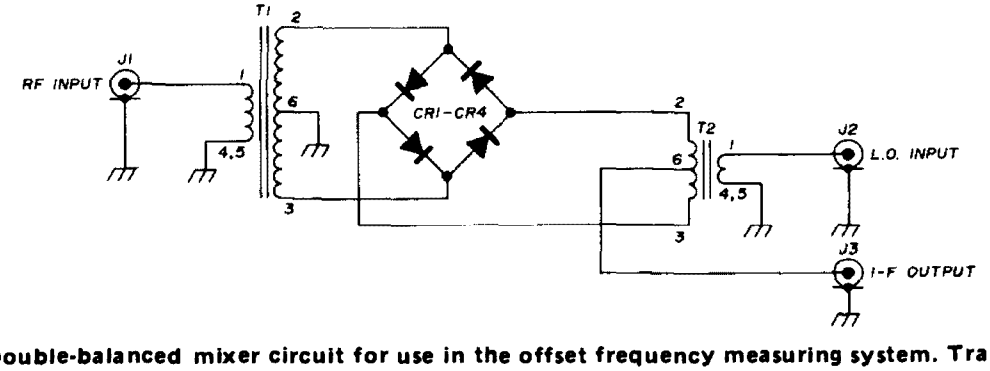


fig. 7. Double-balanced mixer circuit for use in the offset frequency measuring system. Transformers T1 and T2 are Vari-L wideband transformers, model HYB-1. Diodes CR1-CR4 are matched hot-carrier diodes, Hewlett Packard 5082-2805.

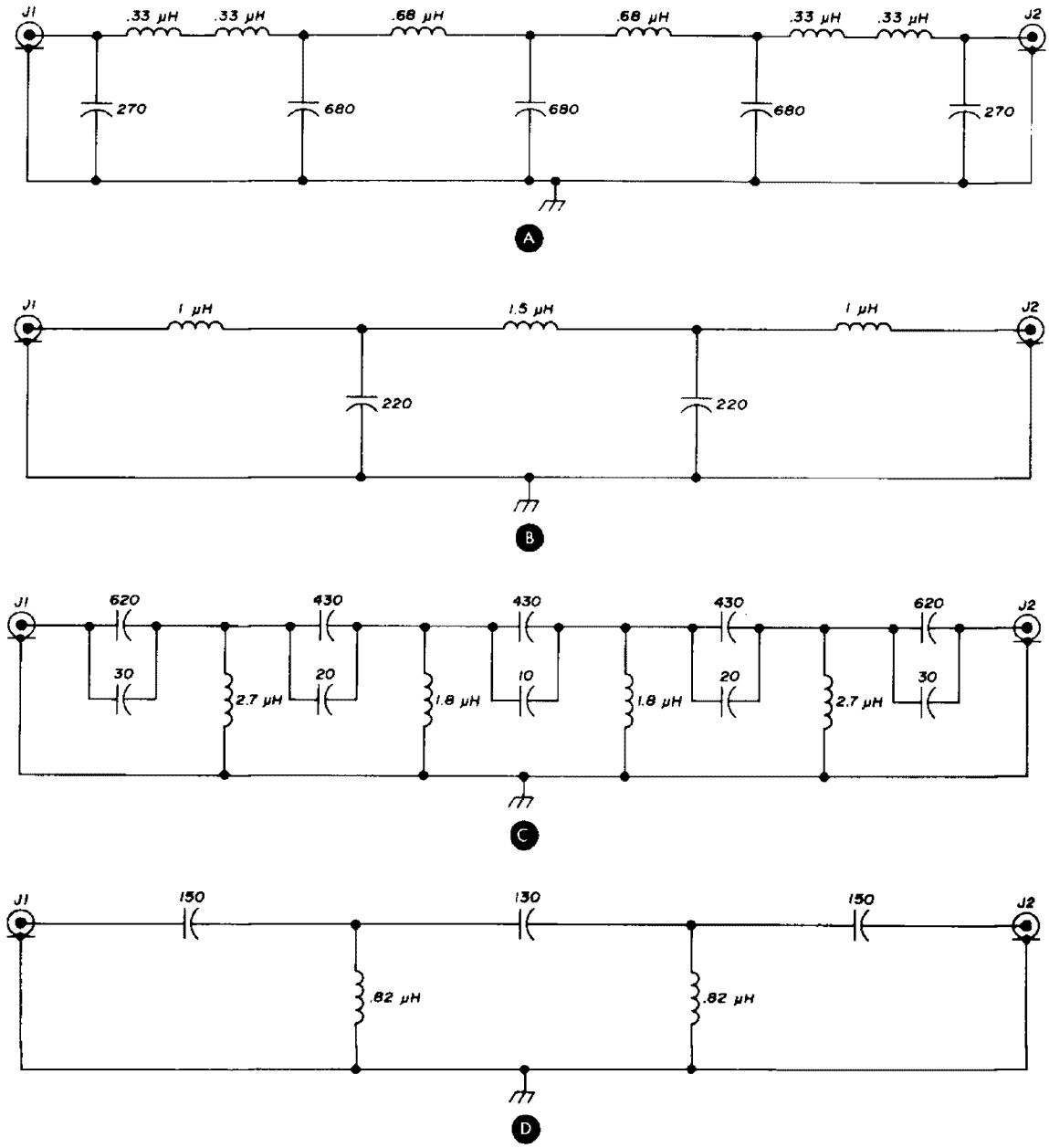


fig. 8. Filters for the offset frequency-measuring system. The filter in (A) is a highpass unit with a 2.9-MHz cutoff (filter A). The highpass filter in (B) has a cutoff frequency of 10 MHz (filter B1). The filter circuit in (C) is a lowpass unit with a 14.2-MHz cutoff frequency (filter B2). The lowpass filter in (C) has a cutoff frequency of 15 MHz (filter C). All capacitors are dipped mica; inductors are Nytronics Wee-ductors.

2.955-3.155-MHz range. The amplifier output is combined in the second mixer with the attenuated 10-MHz output from the time base – the offset frequency.

I elected to take the sum product of the second mixer (12.955-13.155 MHz) for further processing. This frequency range is filtered through filters B1 and B2 in tandem. Filter B1 is a 10-MHz highpass filter and B2 is a 14.2-MHz lowpass design. These two filters may be combined into a single bandpass unit. My junk box dictated the construction of the two filters in tandem. The output of filter B2 is combined in the third mixer with the 75S-3 hfo signal. The difference product is selected by filter C which is a 15-MHz lowpass design.

As indicated, the output of filter C may be fed directly to a counter for hf measurements or on to the vhf portion of the system. Assuming vhf operation, the offset hf signal is fed to the fourth mixer where it is combined with the 62S-1

local-oscillator signal (36-40 MHz for six meters or 130-134 MHz for two meters). Since the 62S-1 local oscillator is on the low side of the received signal, the sum product of the fourth mixer is selected by filter D and subsequently amplified in the vhf amplifier.

Fig. 6 is a block diagram of the 100-kHz counter time base and offset frequency source. It is not anticipated that many people will be interested in duplicating this unit exactly because it includes a surplus 2.5-MHz oven-controlled oscillator. The method and remainder of the circuitry are applicable, however, no matter what oscillator you use.

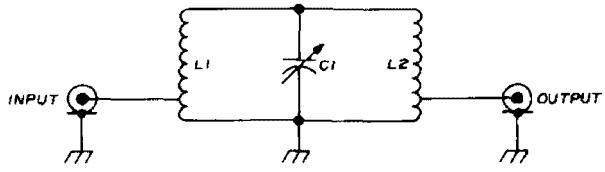
If you have a good 1-MHz oscillator, you could add a times-five multiplier and pick up fig. 6 at point A. If you have a 2.5-MHz oscillator, use the entire block diagram. If you have a 5-MHz oscillator, again enter the diagram at point A. If you have a 10-MHz oscillator, enter the diagram at point B and delete the high-impedance 5-MHz output. If you don't have an existing oscillator available, I recommend a 5-MHz version of the oscillator described by Irv Hoff.[8]

bandswitching

Earlier, I mentioned that bandswitching essentially consisted of changing filters. With the proper selection of filters, more than one band can be covered with the same units. Refer to table 1 for a list of offset frequency ranges for each band when using the 75S-3 receiver and a 10-MHz offset frequency source. It will be recalled that filter C selects the difference product of the third mixer; this is the offset range column in table 1.

If the sum product were selected instead, the lowest frequency of interest would be 19.510 MHz (80 meters), filter C would have to be a highpass type and the hf counter would have to have a much greater operating range. Since 19.510 MHz is the lowest sum product which can cause trouble, filter C should have a cutoff frequency slightly below this value. It can be seen from the offset range column of table 1 (or counted frequency) that if a 15-MHz cutoff lowpass filter is selected for filter C, then all hf bands except 10 meters can be measured with no filter changes.

To measure 10 meters, a 25-MHz cutoff lowpass filter should be substituted at filter C. If you often operate on (and want to measure) 10 meters, you may place both filters in the circuit with a coax



40 - 44 MHz

L1,L2	13 turns no. 16, 5/8" diameter, 1-3/8" long, tapped at 2 turns from ground end
C1	9-180 pF mica trimmer (Elmenco 463)

134 - 138 MHz

L1,L2	7 turns no. 16, 3/8" diameter, 3/4" long, tapped 1-1/4 turns from ground end
C1	7-45 pF mica trimmer (CRL 822BN)

210 - 215 MHz

L1,L2	5 turns no. 16, 3/8" diameter, 1/2" long, tapped 1 turn from ground end
C1	4.5-25 pF mica trimmer (CRL 822AZ)

fig. 9. Schematic of the bandpass filter required for vhf offset-frequency measurements (filter D).

table 1. Offset frequency ranges for each amateur band when using a Collins 75S-3 receiver and a 10-MHz offset frequency source (all frequencies in MHz).

band	hfo	vhf LO	offset range
3.5 - 4.0	6.555 - 6.955	—	6.0 - 6.5
7.0 - 7.3	10.155 - 10.355	—	2.7 - 3.0
14.0 - 14.4	17.155 - 17.355	—	4.0 - 4.4
21.0 - 21.6	24.155 - 24.555	—	11.0 - 11.6
28.0 - 29.7	31.155 - 32.755	—	18.0 - 19.7
50.0 - 54.0	17.155	36 - 40	40.0 - 44.0
144-148	17.155	130 - 134	134.0 - 138.0
220-225	17.155	206 - 211	210.0 - 215.0

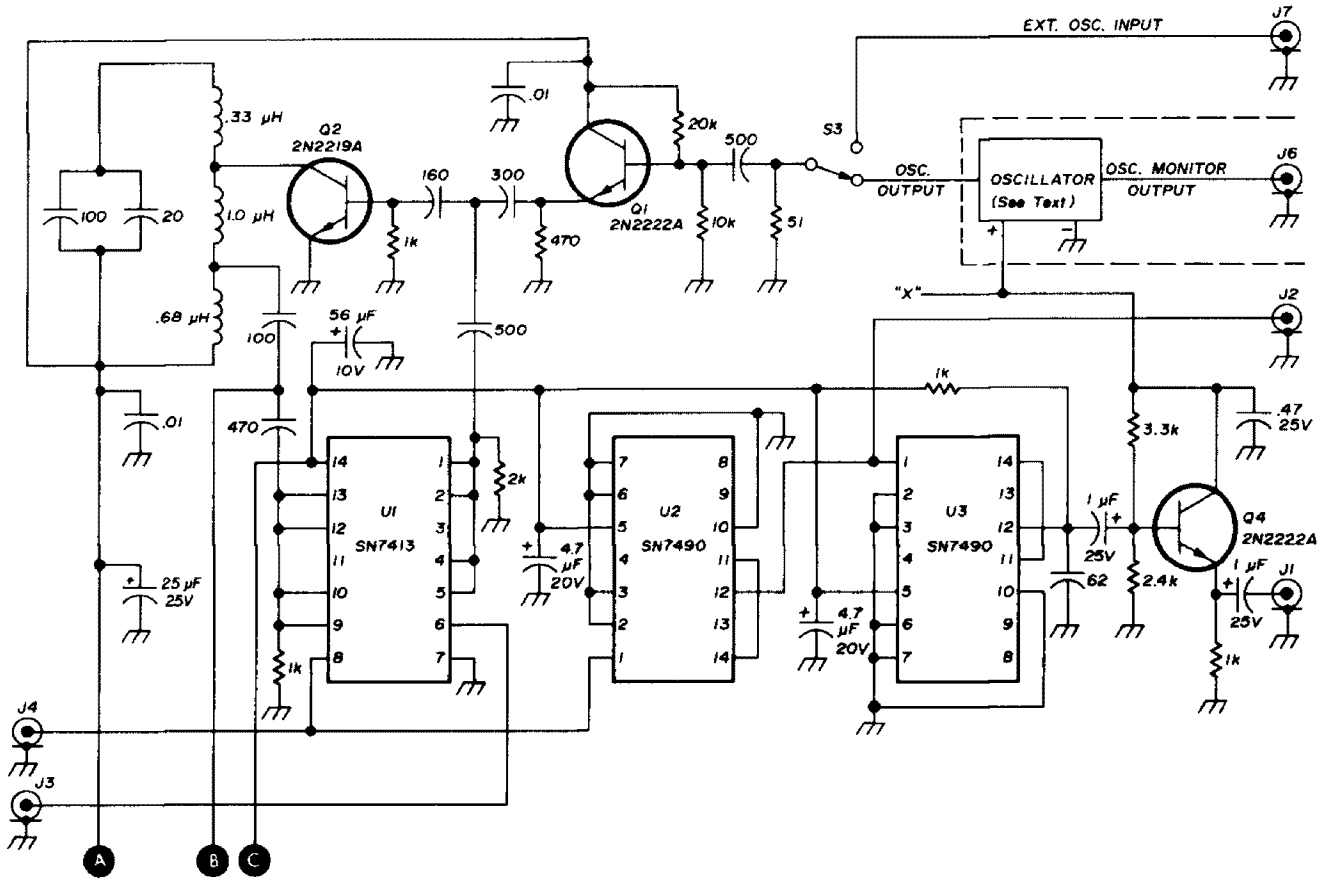


fig. 11. Time-base module used in the offset frequency-measuring system. Transformer T1 is a Stancor P-8180.

switch to short out the 15-MHz filter when operating on 10 meters.

It should be noted that if a 7-, 14- or 21-MHz i-f range is chosen for your vhf converter, then filter C may remain unchanged from hf band operation. Just drive the fourth mixer directly with the output of filter C. Filter D, of course, is selected for the vhf band to be used according to the offset range column of table 1.

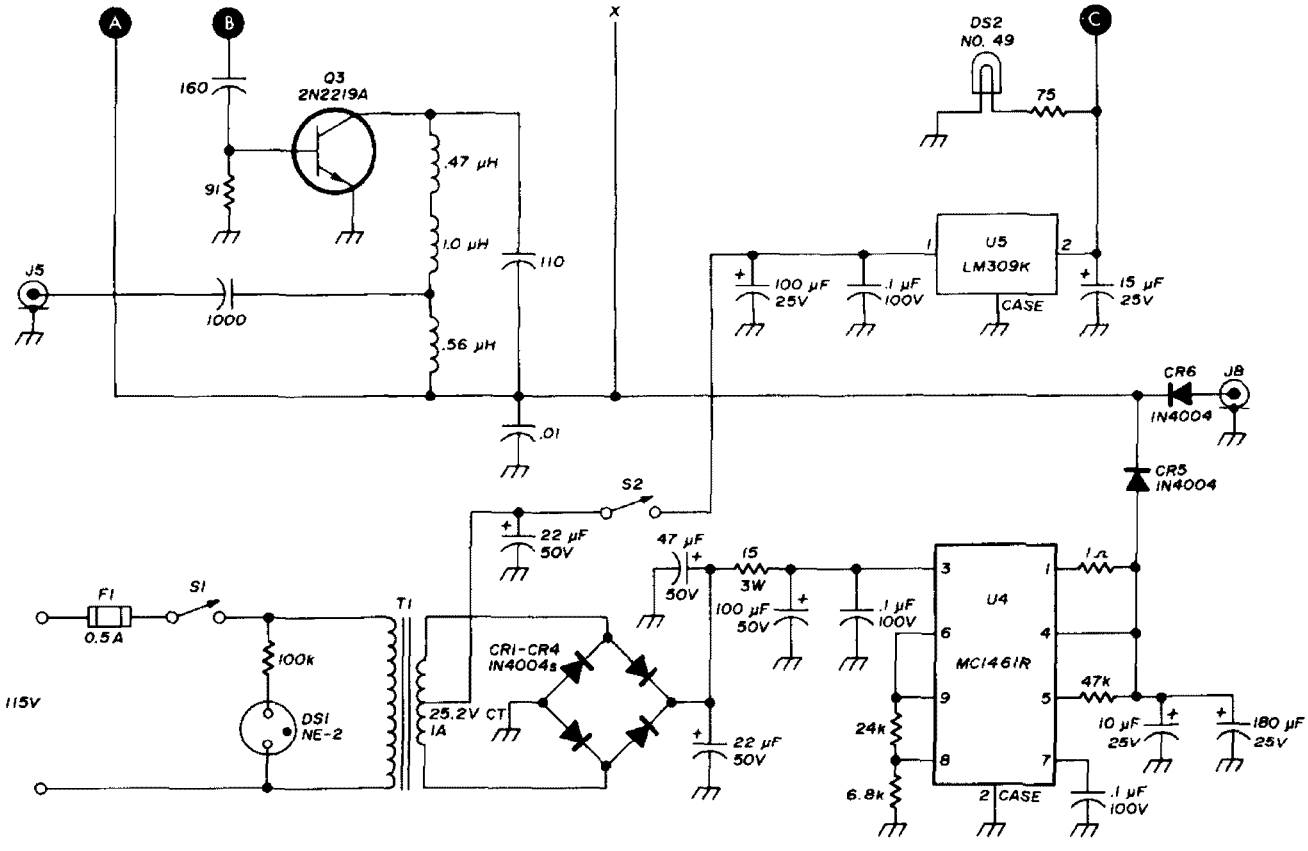
circuit description

The module circuits will be described in the following order; mixers, filter A, filter B1, filter B2, filter C, filter D, 2.955-3.155 MHz amplifier and power supply, vhf amplifier and time base.

All mixers are passive double-balanced mixers using hot-carrier diodes. Double-balanced mixers were chosen because they provide excellent suppression of mixer input signals at the output port. Passive mixers were chosen over active circuits because of their superior performance at vhf. Also, it is easier to provide the higher LO injection levels required by passive mixers than to put together the additional circuitry required by active devices (power supply, input/output coupling, etc.).

Several approaches may be taken to acquire the four mixers required for the system. Double-balanced mixers enclosed in shielded containers with female BNC connectors are commercially available from sources such as Hewlett-Packard, Relcom and Vari-L; prices start at around $40. Mixers with similar specifications but designed for PC board mounting are available from Vari-L, Mini-Circuits Laboratory, Merrimack Industries and others. Prices start at under $10 for quantities greater than five.

If you prefer to build you own mixers, one method has been described by Ress.[9] This unit will cost about $10 plus the PC board. If you can fabricate your own board, this is the most economical approach. I also built a double-balanced mixer using commercial wideband transformers and hotcarrier diodes mounted



on perf-board with adhesive copper foil used as the ground plane. The schematic for this unit is shown in fig. **7.**

Filter A selects the 2.955-3.155 MHz output of the first mixer. This filter must have a sharp cutoff characteristic to attenuate the undesired 2.045-2.245 MHz image. I decided to use a nine section Chebyshev highpass filter with 1-dB passband ripple (fig. **8A**). This type filter has a sharper cutoff characteristic, for a given number of sections, than either the Butterworth or older image parameter designs. It provides a minimum of 16-dB rejection to the image. If your receiver has different i-f or vfo frequencies, this and subsequent filters may be designed using references 10 and 11.

Filter B1 is a 10-MHz highpass design. It is similar to filter A except it has only five sections (fig. **8B**). Filter B2 is a nine-section Chebyshev lowpass design with a 14.2-MHz cutoff frequency (fig. **8C**). Filter C is a five-section Chebyshev low-pass design with a cutoff frequency of 15 MHz (fig. **8D**). Filter D is a lumped-constant bandpass design which yields 3-dB bandwidths in the vicinity of 5% of F_o and ultimate rejection of at least 40 dB (fig. **9**).[12]

wideband amplifier

The 2.955-3.155 MHz amplifier is built into a small Minibox (Bud CU-2100-A). The power supply was built in a separate enclosure for more convenient placement. The amplifier is a single-stage common-emitter design (fig. **10**), only the collector circuit is tuned. The output is tapped from the collector tank capacitor rather than the inductor. This was done so I could use a miniature molded inductor in the circuit. Gain is approximately 25 dB.

The combination of filter D and the vhf amplifier, in fig. 5, provides approximately 20-dB gain at the selected vhf offset frequency. The most versatile way to achieve this is to follow filter D with a wideband 20- to 25-dB gain amplifier having a high-frequency cutoff at least as high as the maximum frequency to be counted. An alternative is to replace filter D and the wideband amplifier with a

multi-stage tuned vhf amplifier. I have used both approaches successfully.

A disadvantage of using the alternative is that every time you go to a different position of a vhf band, you must retune the amplifier or replace it with another pretuned amplifier. If your interest lies in only *one* portion of *one* vhf band, *then* this approach is the most economical. As a matter of fact, if you have a preamplifier for the band(s) of interest, and it will put out 1 milliwatt before gain compression, you can use this in place of filter D and the vhf amplifier.

The preamp must be capable of being tuned to the offset frequency 10 MHz below the band being used. This is most likely possible on 144 and 220 MHz. If you use a broadband preamplifier (10 - 20 MHz) be sure to use filter D to prevent the image from the fourth mixer from getting through.

If you prefer to use the filter and wideband amplifier, commercial models are available from Hewlett-Packard, C-Cor, Avantek, Optimax and others. Prices of modular wideband amplifiers suitable for PC board mounting begin around $40. If you are interested in building your own wideband amplifier, see references 13 and 14 for ideas. Also, the International BAX-1 amplifier module should be useful in this application.

time-base module

The time base is the largest module in the system. Indeed, it could be considered a sub-system (fig. 11). The schematic does not include an oscillator, which may be considered as a separate module. A recommended circuit is given in fig. 12.

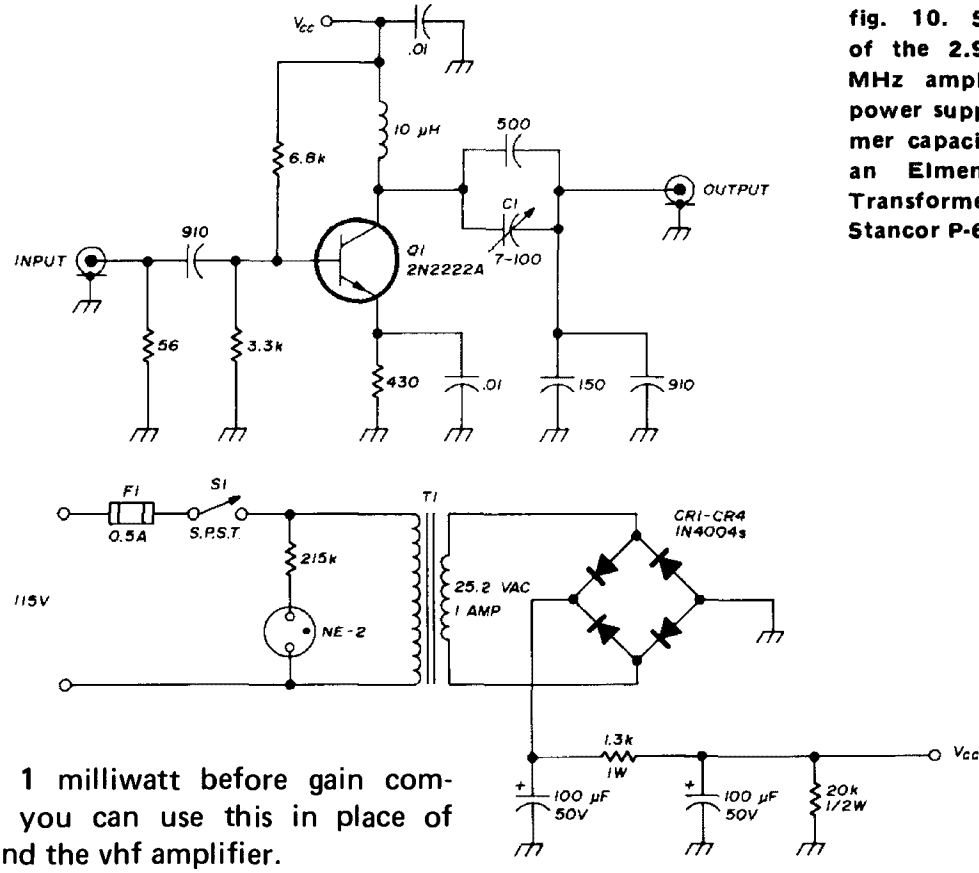


fig. 10. Schematic of the 2.955-3.155 MHz amplifier and power supply. Trimmer capacitor C1 is an Elmenco 423. Transformer T1 is a Stancor P-6469.

Referring to fig. 11, the input stage, Q1, is an emitter follower which establishes a resistive 50-ohm input impedance. This stage drives an SN7413N Schmitt trigger which will give a 2.5 or 5.0-MHz, TTL compatible, output (J3) depending on the oscillator you use. Transistor Q1 also drives a class-C doubler/amplifier stage (Q2) which operates as a doubler when driven at 2.5 MHz and as an amplifier when driven at 5 MHz.

The doubler/amplifier drives the second half of U1 (SN7413N), a Schmitt trigger, which has a TTL compatible 10-MHz output (J4). Transistor Q2 also drives Q3 which is a class-C, 10-MHz output amplifier; Q3 provides +25 dBm (315 mW) output at 10 MHz (J5). A portion of this signal is fed to the external second mixer for the offset frequency.

The 10-MHz output of U1 drives U2, an SN7490N decade divider, which provides a TTL compatible 1-MHz output (J2) and drives U3, an SN7490N decade divider, which drives Q4, an emitter follower. Q4 provides a 100-kHz output (J1) for the counter external time base input.

The power supply uses a center-tapped bridge circuit for dual output voltages. The high output (15 volts) is regulated by an MC1461 IC voltage regulator and the low output (5 volts), by an LM309K regulator. Assuming you install your oscillator on the chassis with the remainder of the time base, provision is made for oscillator power from an external battery/trickle charger to supply standby power during power outages or short term transportation (J8).

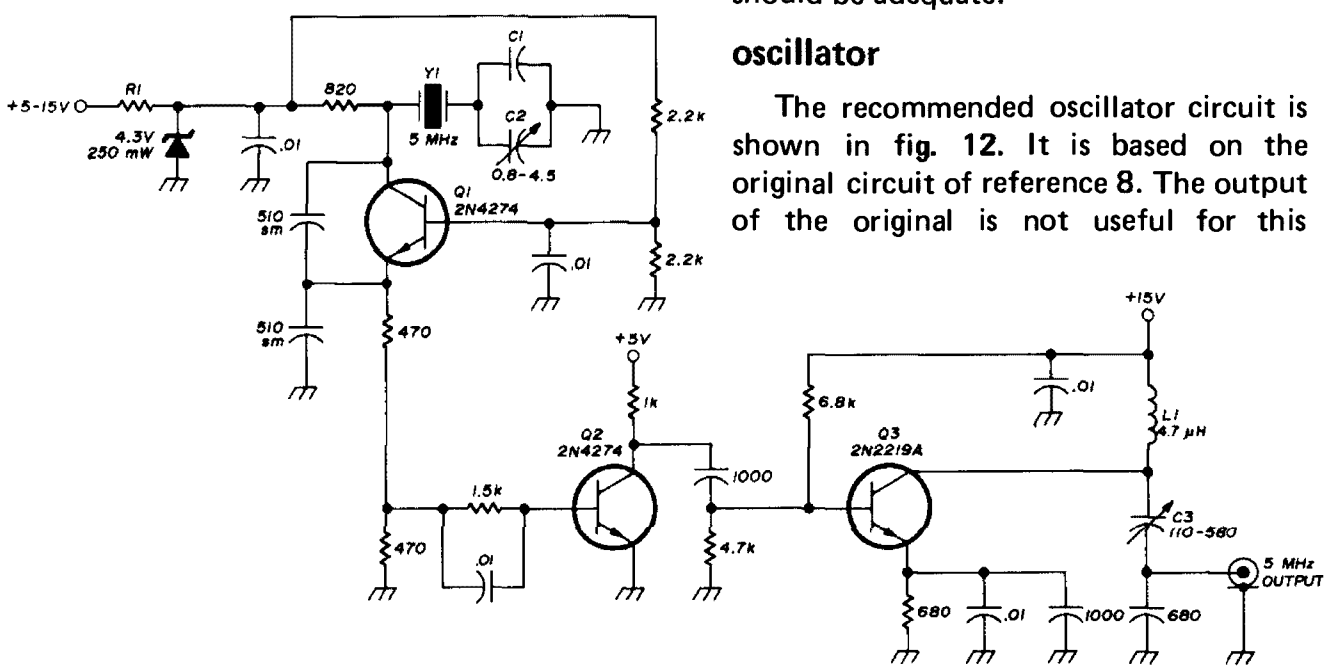
When driving Q1 with an external oscillator (J7), you will need to provide plus 20 dBm (100 mW) of 2.5-MHz signal or plus 15 dBm (30 mW) of 5-MHz signal. No tests have been made to determine how much 10-MHz signal is required to drive Q3/U1 directly but gain data indicate that zero to 10 dBm (1-10 mW) should be adequate.

oscillator

The recommended oscillator circuit is shown in fig. 12. It is based on the original circuit of reference 8. The output of the original is not useful for this application. The output of fig. 12 will drive Q1 of fig. 11 directly. This oscillator circuit may be built on a separate PC or perf board and included on the chassis with the circuitry of fig. 11.

Mount the oscillator on *top* of the chassis, not inside with other heat producing components. Provide as much thermal isolation as possible.

I have not built this oscillator, but the original circuit is well documented and the output stage is similar to the 2.955-3.155 MHz amplifier described earlier. Tuneup procedures for both amplifiers are identical.



C1	Temperature-compensating capacitor	R1	5-volt source, 36 ohms, 1/4 watt; 15-volt source, 510 ohms, 1/2 watt
C2	0.8 - 4.5 pF glass or ceramic trimmer	Y1	5-MHz crystal (International HA-1)
C3	110 - 580 pF trimmer (Elmenco 467)		

fig. 12. Recommended time-base oscillator circuit.

It is possible to duplicate the double-balanced mixer shown schematically in fig. 7 using easy-to-acquire parts. The mechanical details may be seen in the photograph of fig. 13. The enclosure is a Bud CU-2100-A Minibox. The connectors are BNC types (UG-625A/U). The circuitry is built on a piece of perf-board with holes on 0.1-inch centers. The board is cut just over the width of the trans-

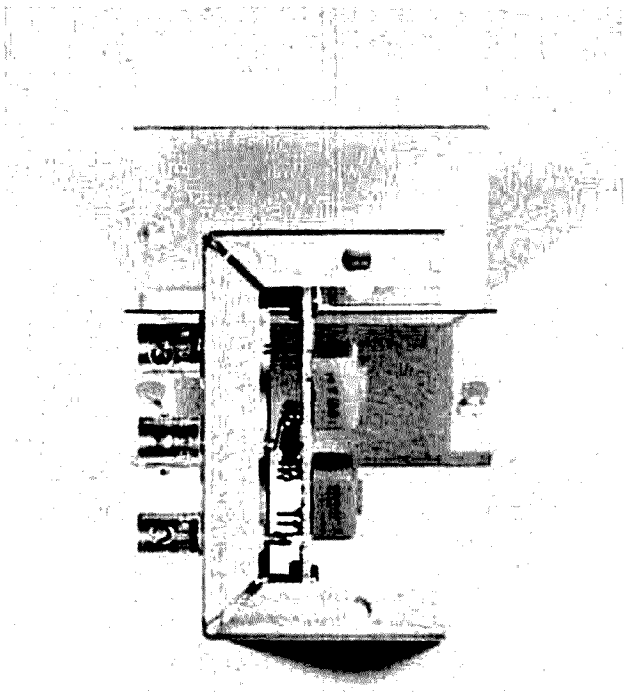


fig. 13. Construction of the double-balanced mixer shown in fig. 7.

formers (0.75 inches) and approximately 1-3/4 inches long. Adhesive copper foil is applied to one side of the board; the two transformers are mounted from the other side.

The two sets of transformer terminals labeled 2, 3 and 6 are placed adjacent to each other on the board and the four diodes are used to interconnect these two windings. The copper foil can be cut away from non-grounded transformer terminals with a razor blade or X-acto knife. Grounded terminals are soldered to the foil with a small 25-watt iron.

The assembled perf-board is mounted diagonally in the Minibox. This allows each of the transformer pin-1 connections to be soldered directly to the center pin of a UG-625A/U connector. Be sure to drill the connector mounting holes on the correct centers for direct insertion of transformer pin-1 terminals. A short piece of bare tinned wire is required to connect the center tap of transformer T2 to the center coaxial jack.

Construction of filters A through C is very similar. A photograph of filter C is shown in fig. 14. The enclosure is a small Minibox; the size depends upon the number of filter sections. The connectors are female BNC chassis-mounting types (UG-625A/U). The filter components are mounted between the connector center pins, small standoff insulators (about 1/2-inch high) and strategically located ground lugs. Use dipped mica capacitors and Nytronics Wee-Ductors or equivalent for the inductors. Keep individual component leads short and the total circuit path as direct as possible. Where possible, arrange the inductors at right angles to one another for minimum mutual coupling.

Filter D is constructed using the same type of enclosure and connectors as the previous filters although the circuit elements are arranged differently. The general layout is shown in the photograph of fig. 15. This is the 134-MHz filter. Half-inch standoff insulators are used to support the inductors and are mounted as close to right angles to one another as possible.

The 2.955-3.155 MHz amplifier is also built into a Bud CU-2100-A Minibox. The circuitry is built on a small piece of perf-board with holes on 0.2 inch centers. The components are inserted from one side and the leads soldered together on the other side. The layout is shown in fig. 16. The board is mounted on 1/2-inch spacers. The input, output and power connectors are UG-625A/U. The power supply for the amplifier is built in a separate Minibox which is large enough to contain the components. No particular construction technique is required in this unit – just keep it small and neat.

The time base is built on a 5x7x2-inch aluminum chassis. The switches, connectors, pilot lights and voltage regulators are mounted on the front, rear and side panels of the chassis. The oscillator is mounted on the top surface and the circuit boards are mounted inside. Fig. 17 is a photograph of the unit.

The rf and digital circuitry of fig. 11 is built on the main board plus a small piggy-back board. The power supply and control wiring is accomplished in a point to point manner within the chassis. The main board contains the analog circuitry (Q1, Q2, Q3 and associated components) lengthwise on one side and the digital circuitry on the other. Q4 and its associated components are on the small board.

Outputs from the boards are via push-in terminals and attached sections of miniature RG-174/U coaxial cable. All board-mounted components are inserted from one side and leads soldered on the opposite side. The large board is mounted on the chassis top surface using two half inch spacers. The small board is mounted on a one-inch spacer screwed to the main board. Keep the coaxial cables carrying the output signals as short as possible. This is particularly true of the 5- and 10-MHz high impedance outputs.

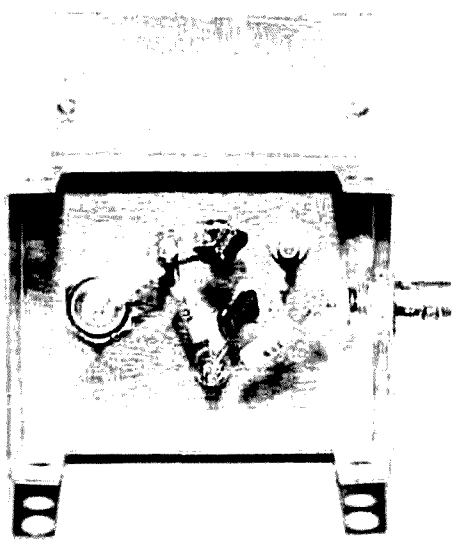


fig. 14. Lowpass filter with 15-MHz cutoff (filter C).

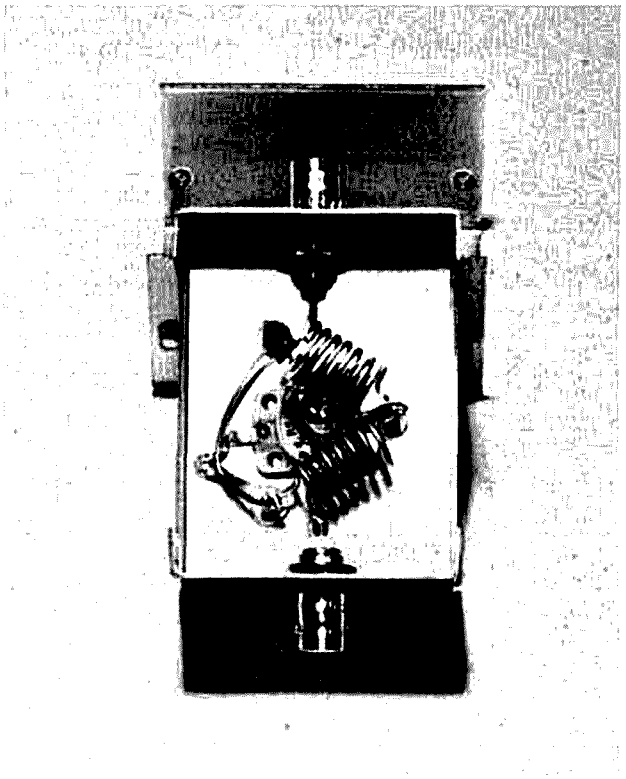


fig. 15. Construction of the 130-MHz bandpass filter (filter D).

If the oscillator is mounted on the chassis top surface, the terminals should protrude into the chassis interior for power and rf output connections. When wiring the power supply portion, use point-to-point wiring. It is important that all component leads be kept to a minimum length. This is particularly true of the MC1461 which has the larger number of external components and contains active devices capable of sustaining high-frequency oscillations. Use the same wiring precautions with both voltage regulators that you would use with any high-frequency rf circut. The parts location and wiring of the remainder of the power supply is not cricical.

adjustment

The i-f vfo and hfo signals must be tapped and brought out to the measurement system. I used the technique suggested in reference 1 but component values were revised slightly as shown in fig. 18. It may be necessary to vary the value of the 45-pF capacitor across L1 to maximize the i-f output level. With this

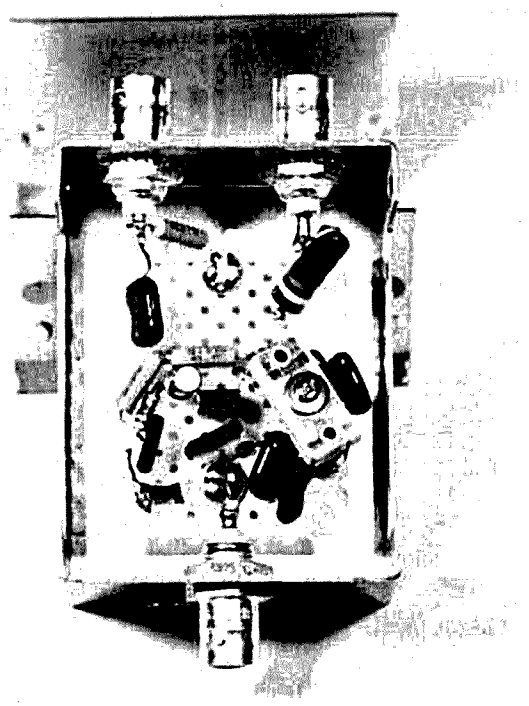


fig. 16. Layout of the 2.955-3.155 MHz amplifier module.

method of obtaining signals for the frequency measurement system I have observed no degradation to normal 75S-3 performance. If you desire further isolation for your receiver, you may use mosfet amplifiers such as used in the Heath SB-650.[5]

The vhf local-oscillator injection voltage is obtained from the 62S-1 by using a homebrew tee adaptor at the input to the 148-MHz trap (J31 of FL2 in the 62S-1). The tee adaptor is made from two standard phono pin connectors. The male portion is a phono plug to miniature phone jack adaptor (Lafayette 99R63455 or Switchcraft 365). To gain access to the center pin terminal, remove and discard the shell. The female portion of the adaptor is a single hole mounting phono jack (Lafayette 99R62341). The center pins of the two connectors are soldered together. The ground lug of the female connector is then connected to the remaining portion of the shell of the male connector.

The coaxial cable running to the frequency measuring system is then soldered (center conductor and shield) to the adaptor at the junction of the two component connectors. The male end is then inserted in J31 of FL2 and the cable which originally plugged into J31 now plugs into the female portion of the adaptor. The RG-174/U coaxial cable from the tee adaptor is then run *through* the center hole of unused phono jack (J13) on the rear apron of the 62S-1. Thus, no permanent change to the transverter is necessary.

Some experimentation with the length of the vhf local-oscillator cable and how it is terminated may be required. My cable is 41 inches long; if I leave the cable unterminated, a fairly low impedance is reflected back to the tee adaptor. Receiver and transmitter LO injection is thus reduced and vhf receiver performance and transmitter output power are affected.

If the cable is terminated in 50 ohms or lengthened about 12 inches and left unterminated, 62S-1 receiver and transmitter performance are normal. This effect could be completely eliminated by using a broadband 36 to 134-MHz amplifier with a high input impedance at the tee adaptor. Such an amplifier would have to be tailor fitted to the 62S-1. However, it is reasonably easy to obtain normal 62S-1 operation without resorting to a buffer amplifier.

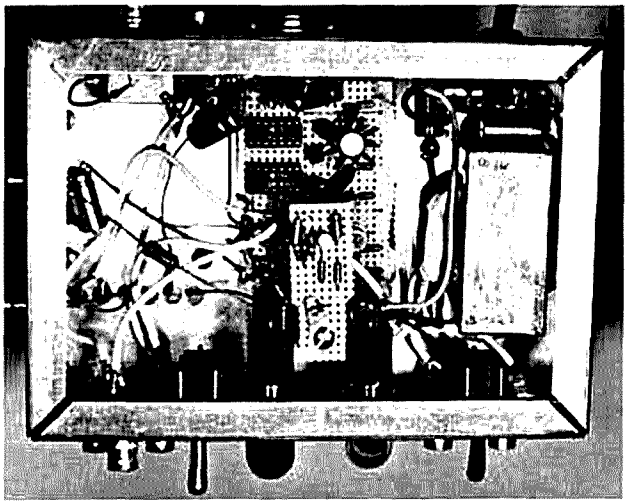


fig. 17. Bottom view of the time-base module. The MC1461 voltage-regulator IC and BNC connectors are mounted on the side of the chassis.

Using the circuit connections described here, you should get the power levels shown in table 2 at the output end to the respective cables.

Filters A through C should not require any adjustment after construction. However, to insure that there are no faulty components, it would be wise to check the cutoff frequency and ultimate rejection of each filter. This will require a signal generator (with frequency calibration) and a volt-meter (or powermeter) with a frequency response higher than the highest filter cutoff frequency (15 MHz).

Filter D must be tuned-up. This will require a calibrated signal generator and voltmeter or powermeter with adequate frequency range (220 MHz). A swept-frequency generator plus detector would be better. In any event, C1 sets the filter center frequency and the tap points on L1 and L2 determine bandwidth and insertion loss. The taps given in fig. 9 should be satisfactory. Capacitor C1 should be tuned to the offset frequency region of interest. If you want to measure frequencies more than a few hundred kilohertz apart, you should re-peak C1 for each different region.

table 2. Signal levels available from the Collins 75S-3 receiver and 62S-1 transverter.

signal	condition	level (Into 50 ohms)
I-F	S9 signal	−5 dBm
VFO	—	−7 dBm
HFO	80 and 40 meters	+3 dBm
HFO	20 and 15 meters	−7 dBm
HFO	10 meters	−10 dBm
VHF LO	—	+5 dBm

Although the mixers *have no* adjustment provisions you should insert a known rf and LO signal and check for proper mixing action and suppression of undesired feedthrough products. Feedthrough from rf and LO ports to the i-f port should be more than 20-dB down, referenced to the input levels at the rf and LO ports, respectively. Performance less than this may indicate a bad or mismatched set of diodes.

To properly tuneup the 2.955-3.155 MHz amplifier, a calibrated signal generator and voltmeter or rf powermeter operating to beyond 3 MHz are required. Set the signal generator to 3.055 MHz and adjust C1 for maximum gain (about 25 dB). To optimize the combination of filter A and the 2.955-3.155 MHz amplifier, connect the two in tandem and run response measurements at 2.955, 3.055 and 3.155 MHz. It should be possible to skew the response curve of the amplifier down in frequency to make the overall gain at 2.955 MHz more nearly equal to that at 3.155 MHz (on the low side you have filter *and* amplifier rolloff, while on the high side you have only amplifier rolloff).

Your vhf or broadband amplifier plus filter D should be initially adjusted for 20-25 dB overall gain at the offset frequency to be measured. For example, if you are measuring frequencies near the low end of two meters, the vhf amplifier or filter should be tuned to 134 MHz. Ultimate gain adjustment will depend on counter sensitivity.

The only adjustments necessary to the time-base module will be to the analog section. The tank circuits for Q2 and Q3 may be optimized by starting with slightly lower value capacitors and increasing the value with small padders until the respective outputs peak. Varying the value of the 62-pF capacitor from pin 12 of U3 to ground may result in a cleaner 100-kHz output waveform.

The value of the resistor from pins 8 to 9 of the MC1461 voltage regulator IC determines the high output voltage. A value of 24k ohms results in about 15 *volts out*. This value is not sacred; If you want to change the voltage for your oscillator, it may be done. Lowering the voltage will result in less 10-MHz (low impedance) signal output.

If you increase the value be careful not to exceed transistor breakdown voltage. To increase the voltage substantially, you will need a different power transformer. The maximum current drain for the MC1461 regulator is 500 mA; be sure

your oscillator load does not exceed this value. If you use a non-oven oscillator there should be no problem.

operation

Typical offset counter readings for specific received frequencies on each band are shown in table 3. It should be evident from this table that on all bands except 80 and 40 meters you simply add

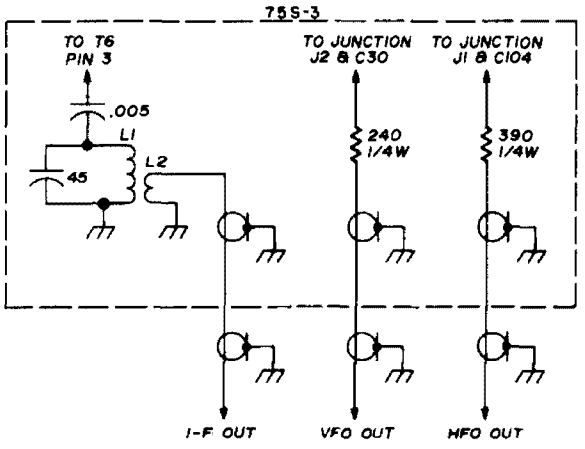


fig. 18. How to connect the Collins 75S-3 receiver into the offset frequency-measuring system. L1 is a 2.5-mH rf choke (miller 6302). L2 is 5-1/2 turns no. 26 wound between two end pi sections of L1.

10.000 MHz to the counted frequency to arrive at the actual received frequency. On 80 and 40 meters you subtract the counted frequency from 10.000 MHz to obtain the received frequency.

80 and 40 meters can be offset in a manner so that to obtain the received frequency you simply subtract 10.000 MHz from the counted frequency. This is accomplished by making the second mixer (fig. 5) a difference mixer and the third mixer a summing mixer. However, there are two disadvantages to this scheme. First, you have to change or bandswitch new filters at B1, B2 and C; second, you cannot count 80 and 40 meters on a 10-MHz counter. For these reasons, I did not use this arrangement.

If a spectrum analyzer or frequency-sensitive voltmeter is available when you are initially assembling the modules into the system, it may be enlightening to observe the output spectrum of each mixer/filter combination. The undesired mixer products should all be down approximately 20 dB. If they are not, the individual mixers and/or filters should be checked for proper operation.

When installing the mixers, be aware that there is an rf port (R), LO port (L) and an i-f port (I). I have achieved best conversion loss and isolation performance in this up-conversion application by applying the lowest frequency to be mixed to the i-f port and taking the up-converted output from the rf port.

In fig. 5 no amplification is shown following filter C for HF operation. If your counter doesn't have a built-in preamplifier, you may need to include one or more stages of gain at this point. The amount of gain and whether it be provided by tuned or wideband stages will be determined by things such as counter sensitivity, amplifiers presently available to you and the state of your junk box and/or pocketbook.

System calibration may be achieved by setting your time-base oscillator against a laboratory standard of known frequency or by using WWV. To use WWV with the 75S-3, tune the receiver to 15 MHz and during a period of no tone modulation, place the receiver in the CW mode (narrowest bandwidth) and center the WWV carrier in the passband. Your counter should now read 5.000 MHz plus/minus 1 count offset frequency. If it does not, adjust your time base oscillator until it does. If the counted frequency is not even remotely close to 5.000 MHz, do not adjust the oscillator, something else is wrong and this must be corrected first.

table 3. Typical offset frequencies for each band (all in MHz).

band	received frequency	offset frequency (counted)
3.5	3.774368	6.225632
7.0	7.024732	2.975268
14.0	14.251051	4.251051
21.0	21.246188	11.246188
28.0	29.499987	19.499987
50.0	50.101687	40.101687
144	146.160752	136.160752
220	222.561357	212.561357

Be sure that you make this adjustment when propagation conditions between WWV and your location are quiet. Typically, this will be during a daylight period at both WWV and your location. If the WWV signal is varying widely in strength or is of marginal strength, postpone the adjustment.

System stability on all bands should be excellent. No whistles, howls or low-frequency motorboating should be heard with the receiver operating normally and the measuring system activated. The ultimate test is to have the normal station antenna connected to the receiver and the counter on and counting. If your system is stable under these conditions on all bands, you are in business. If not, suspect mixer leakage as the primary culprit.

Leakage of 2.955-3.155 MHz energy through the second mixer, filters B1 and B2 and the third mixer is the main sneak path. Another possibility is radiation from the 2.955-3.155 MHz amplifier to the third mixer. If you have one mixer with particularly good rf/i-f port isolation, use it for the second mixer.

summary

Results at my station have been most rewarding. Previous systems based on signal synthesis proved to be oscillatory, very difficult to tune up and always in need of tweaking. The present offset system provides the capability of making frequency measurements on incoming hf signals at least 20 dB weaker than with my best performing signal-synthesis scheme. On two meters, system sensitivity has been increased by more than 40 dB using the offset technique.

Accuracy in the ARRL Frequency Measuring Tests (relative to umpire measurements) has averaged better than 0.2 parts per million for ten measurements submitted during the last 20 months. Most of these measurements have been on the lower amateur frequencies (80 meters) which results in the average absolute error being less than 1 Hz. A histogram of my system error relative to ARRL FMT umpire measurements is shown in fig. 19.

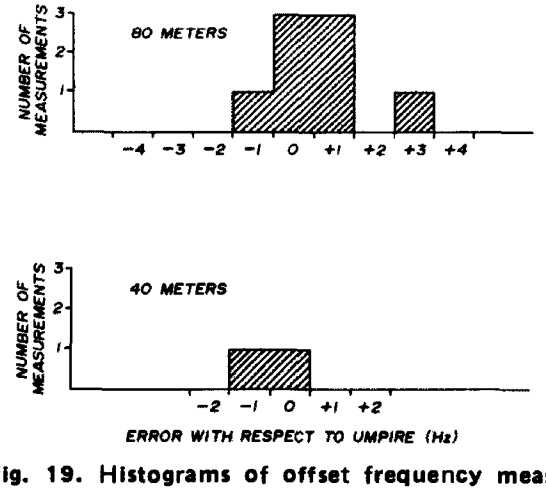
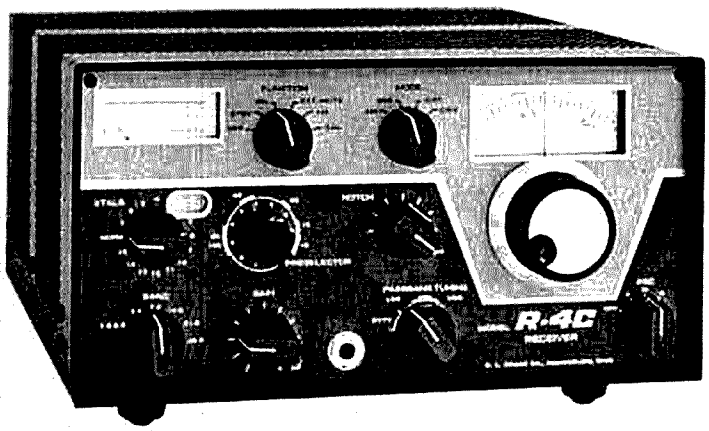


fig. 19. Histograms of offset frequency measuring system error vs umpire readings in recent ARRL Frequency Measuring Tests.

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ham radio



electronic bandpass tuning in the new Drake R4C

The new Drake R4C amateur receiver uses a long-neglected technique to provide a simple system for electronic bandpass tuning

D.H. Horner, 186 Golfwood Drive, Dayton, Ohio 45449

Many amateur operators believe bandpass tuning is an invaluable adjunct to the main communications receiver in their station. It enables one to adjust the exact bandpass of the receiver in relationship to the bfo frequency so that interference from other stations can be dodged, another operator's voice characteristics can be made more pleasant, or a pleasing pitch can be obtained from CW. Also, the proper bandpass can be set for RTTY reception.

Many years ago the Collins Radio Company incorporated bandpass tuning in their 75A-4 receiver by rotating the PTO assembly at the same time as the bfo capacitor was tuned. This gave the advantage of bandpass tuning along with the flat-topped and steep skirt selectivity afforded by the Collins mechanical filter. This was accomplished mechanically, of course, and worked well because Collins maintained extremely close manufacturing tolerances; zero-beat moved no more than 50 Hz during the adjusting process.

The R.L. Drake Company has featured bandpass tuning in many of its amateur communications receivers by tuning the resonant frequency of a four-pole LC

filter operating at 50 kHz. The bfo frequency remains fixed, and the actual i-f is tuned to one side or the other by an amount selected by the operator. This system has proved to be very successful, and is readily attested to by the continued popularity of the Drake receivers.

There is a small drawback, however. With the advent of mechanical and crystal filters, 6-dB to 60-dB skirt ratios in the order of 2:1 or even 1.7:1 are now expected by the amateur fraternity, with the ultimate stopband rejection in the order of 80 dB or so. The 50-kHz LC filters do not give this magnitude of performance.

Right after World War 2 a large segment of amateurs purchased sophisticated surplus communications gear at a fraction of the original government cost. Among the units available was the venerable BC-312, a military communications receiver using design techniques not followed today. This set used a last i-f of 915 kHz, and was rather broad. When ssb became popular, some sort of sideband selection became necessary. Some clever amateurs devised a system where the bfo in the BC-312 was turned into an injection oscillator, a mixer stage was added, and the i-f was mixed down to some lower frequency.

Some builders made their own 50 kHz i-f amplifiers; others cannibalized 85-kHz i-f transformers from the ubiquitous BC-453 command set. Then — and this is the crux of the system — the same bfo was used to mix back to 915 kHz where an added oscillator and detector operated. By varying the frequency of the vfo, which now operated at a frequency removed far enough from 915 kHz to mix to the new i-f frequency, the signal could be made to sweep across the low frequency i-f, and bandpass tuning was the result.

To the extent of my knowledge this technique remained in limbo until the R.L. Drake Company announced the new R-4C Amateur Receiver in March, 1973. This receiver uses electronic bandpass tuning coupled with the selectivity afforded by 8-pole crystal bandpass filters. However, the mixing scheme is different from that used with the BC-312 (see fig. 1).

The rf signal is first converted to 5645

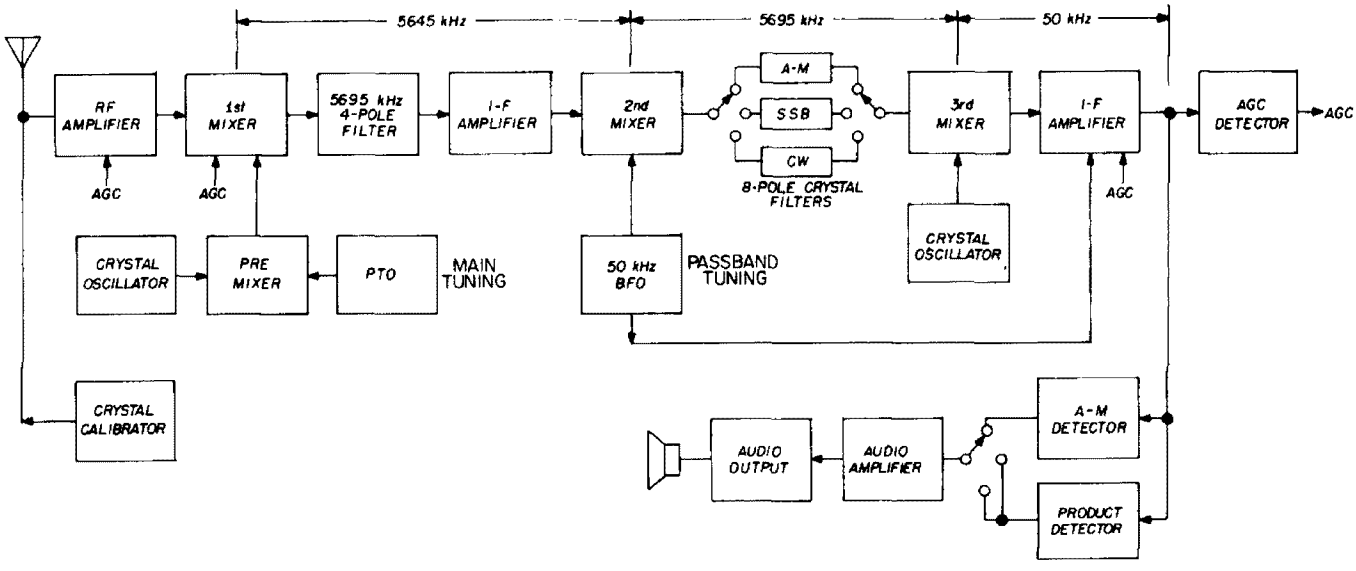
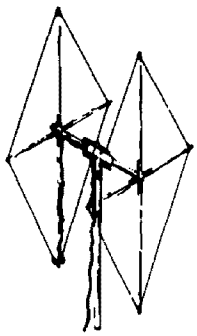


fig. 1. Basic block diagram of the new Drake R-4C communications receiver which features electronic bandpass tuning.



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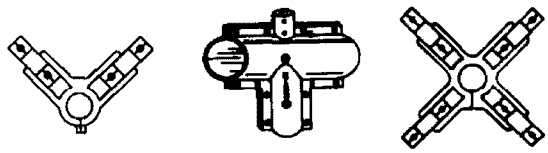
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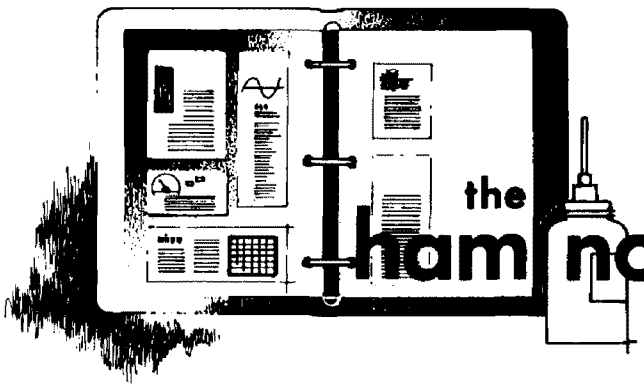
kHz where a 4-pole crystal-lattice filter does much to ward off the degrading effects of strong stations near the operating frequency. Converting all bands into the high i-f is accomplished by mixing the output of the vfo with various crystal-controlled frequencies so the i-f is the same for all bands. Then, a tunable oscillator at 50 kHz is used to up-convert to 5695 kHz. The standard 2.4-kHz ssb filter as well as the accessory filters for CW and RTTY operate at this frequency. Then, the 5695-kHz signal is mixed with a 5645-kHz crystal oscillator to obtain a 50-kHz output. Two tuned LC circuits at 50-kHz help establish "distributed selectivity."

The same tunable oscillator that is used to mix from 5645 kHz to 5695 kHz also serves as the bfo. Because it both up-converts and down-converts (i-f to audio) at the same time and by a similar amount, zero-beat does not vary during adjustment of the bandpass tuning control.

The Drake R-4C Communications Receiver also has provision for a-m. However, as bandpass tuning is not needed in this mode, the additional conversion process is not used. Filters with 4.0- and 6.0-kHz selectivity are stocked by the Drake Company, but these are at 5645 kHz, rather than at 5695 kHz where the ssb and CW filters operate. In the a-m mode a different crystal frequency (5595 kHz) is used to convert to the 50 kHz i-f.

This dual-conversion electronic bandpass tuning technique is simple and relatively inexpensive for the manufacturer to build, and it makes operating much more pleasurable than with receivers which have fixed bfo/filter frequency relationships. One benefit of this particular bandpass tuning technique is less noise in the audio output – receivers that lump all the selectivity in the filters and then use broadband i-f amplifiers following them tend to have "broadband hiss" or white noise appear at the output. The Drake R-4C is pleasantly free from this effect.

ham radio



the ham notebook

portable fluorescent light

Have you ever wanted one of those portable fluorescent lights to use during field day or on camping trips? They are quite expensive and are not used too often, but they're handy to have around. If you have a well supplied junkbox and a little ingenuity, you can build your own.

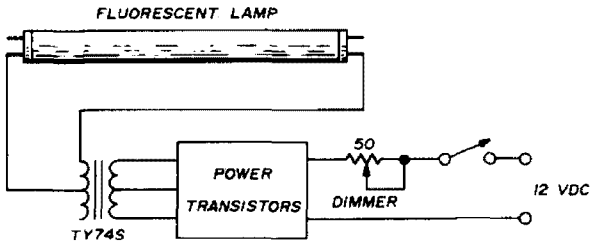


fig. 1. Portable fluorescent light uses transistor power transformer cannibalized from an old CB set. Vibrator power supply can also be used as described in the text.

Most amateurs have some old portable vacuum-tube equipment or CB rigs laying around that can provide most of the parts you need. Try to find a Triad TY74S transistor power transformer or something similar. This will provide the high-voltage supply for your portable fluorescent light.

Remove the power transformer and the power transistors from the old rig, but discard the rectifier diodes and the filter network. You won't need them. Build a metal or wooden framework similar to that found on commercial portable fluorescent lights so you can mount the fluorescent lamp with sockets. The switch, transformer and power transistors can be mounted on the back or in the base.

When you're putting the unit together be sure to use a heatsink. If the light is used continuously there will be a fair amount of heat generated. However, the heat sink doesn't have to be anything fancy, a small aluminum chassis will work.

Wire up the unit as shown in fig. 1 using the original transistor oscillator circuit. Connect the center tap of the power transformer to one end of the fluorescent lamp and one of the transformer end windings to the other end of the lamp. If you want to include a dimmer control place a 10-ohm, 50-watt rheostat in the primary lead as shown in the schematic. My unit required from 0.5 to 1.6 amps, depending upon the setting of the dimmer control.

If you can't find an old high-voltage transistor power supply, you can accomplish the same thing with an old vibrator power supply. These are plentiful – just get an *old* car radio from your local junk yard. Disconnect the filter network and bring out the center tap from the vibrator transformer to one side of the fluorescent tube. Run the other lead to the other end of the tube as shown in fig. 1.

Ken Gray, K8BYO

great circle charts

A number of Great Circle Distance and Azimuth Charts are available from United States Government sources. These may be obtained from numerous local sales agents and distribution centers, or by

mail from the U.S. Naval Oceanographic Office, Washington, D.C. 20390.

Such charts are available at $1.50 centered as follows: 5180, Fairbanks, Alaska; 5181, Seattle; 5182, Honolulu; 5183, Guam; 5184, San Francisco; 5185, Washington, D.C.; 5186, Moscow; 5187, Adak, Alaska; 5188, Kodiak, Alaska; 5189, Eniwetok Atoll; 5190, San Diego, Calif.; 5191, Cutler, Maine; 5192, Balboa, Canal Zone; 5193, Yosami, Japan; 5194, Australia-N.W. Cape; and 5195, Keflavik, Iceland.

For those purchasing a computer readout, possibly they can plot the results conveniently on a 5142 Azimuthal Equidistant Projection of a Hemisphere, priced at 50 cents.

Bill Conklin, K6KA

using the HW-101 transceiver with a separate receiver

Modifications to transceivers such as the Heath HW 100/101 or Swan 350 for split frequency operation have been described as has incremental tuning by means of a variable capacitance diode[1] or operation in conjunction with a separate vfo.[2] In both cases some limitations remain; cross-band operation is not possible and the separation between receive and transmit frequency must remain small because the driver and receiver preselector circuits cannot be independently peaked.

Conversion for use with a separate receiver, using the built-in TR relay, is simple and eliminates these limitations. It is described here for the case of the HW-101.

Two rf-quality phono jacks are installed on the rear chassis apron into the compartment containing the antenna connector, located just below the rf cage. The coax lead from the rf driver circuit board (the receiver input) is disconnected from lug 4 of the relay and from the adjacent ground lug. If sufficient in length, this coax is connected directly to the nearer of the two jacks just installed. Otherwise, it is first extended or replaced by a longer section. Another 50-ohm coax lead is connected to the more distant of the two jacks at one end and to lug 4 of the relay (inside conductor) and to the adjacent ground at the other end. The trap coil (L905) is not disturbed.

In the transceive mode an external coax connector (50-ohm with two rf-quality phono plugs) is jumpered between the two new jacks; the transceiver then functions as before the modification. For separate receive operation a 50-ohm coax lead is connected between the receiver antenna input and the jack, which is fed from lug 4 of the relay. The other jack remains unconnected. To spot the transmit frequency on the receiver the mike/CW level control is turned counterclockwise below the point where the relative power meter indicates any measurable output; the transmitter is then keyed. This provides sufficient signal for spotting without radiating an appreciable amount of power. All that is needed to transmit is to advance the mike/CW level control to the point of full rf output.

This simple modification leaves the transceiver circuitry essentially undisturbed and permits semi break-in operation, cross-band or in-band operation with any desired frequency separation between receiver and transmitter, with both tuned for peak performance.

Incidentally, the modification permits the insertion of an rf-preamplifier, tuned or broadband, between the output coming from the TR relay and the internal or external receiver. A low-noise preamplifier improves the signal-to-noise ratio, especially on 10 and 15 meters.

references

1. P. Pagel, K1KXA, "Receiver Incremental Tuning for the Swan 350," *ham radio,* July, 1971, page 64.
2. R.J. Bacher, "Split-Frequency Operation With the Heathkit HW-100," *CQ,* May, 1970, page 32.

Max Blumer, WA1MKP



comments

Ic speech clipper

Dear HR:

Although K6HTM seems satisfied with the performance of his speech clipper, his conclusion that it's "not much better than a typical agc-type compressor" caused me to take a critical look at the schematic. The main reason for what I would regard as poor performance is fairly obvious. The output impedance of pin 9 of the IC, to which the clipping diodes are connected, is that of an emitter follower and given as 85 ohms on the data sheet.

If author Bird uses 1N75 diodes for the clipper he would obtain virtually no clipping since their ac resistance is likely to be several hundreds of ohms. Even with hot-carrier diodes, clipping will not exceed 4 dB or so. This would definitely make the device perform comparable to alc or agc schemes! A resistor (2000 ohms for hot-carrier diodes, 20k for 1N75s) in series with the 0.01-μF coupling capacitor will improve matters.

I am distressed at the author's lack of concern for carrier rejection. Unless the peak carrier level, at pin 9 of the LM373, is appreciably lower (at least 10 dB), than the voltage at which clipping starts, the advantages of rf clipping will be lost. K6HTM probably forgot to mention the need for precise frequency adjustment of the crystal oscillator, so that the proper low frequency response is achieved, as well as some additional carrier rejection.

Some builders may experience difficulty, and possibly instability, due to undesired coupling via the two oscillator injection points, which are, in fact, connected together. Fortunately, the level at pin 6 of the LM373 is much higher than needed (by a factor of 100), so the desired isolation can be obtained by means of a resistive attenuator.

The second mechanical filter, FL2, is quite superfluous. Filtering after the clipper is needed to prevent rf harmonics from reaching the final mixer. A moderate Q pi network is quite adequate for this purpose and obviously much less costly. As the available post clipper gain is much larger than needed, two 10,000-ohm resistors, one in series with "high" side of the output level control, the other in series with the FL2 input, will give desirable isolation of the filter from the clipping diodes.

If Mr. Bird will modify his device in accordance with this letter, he will obtain results similar to the ones I get with the Comdel unit.

Walter Schreuer, K1YZW
Ipswich, Massachusetts

I am pleased that Walter Schreuer read my article, "IC Speech Clipper" in the February, 1973, issue of ham radio. *What follows is a reevaluation encouraged by his suggestions and some additional information.*

The 1N75 diodes first used in this project were abandoned early in favor of hot-carrier diodes in both the balanced modulator and clipper sections. Carrier balancing with these devices far exceeds the 10-dB figure referred to by Mr. Schreuer, and the total suppression after the first filter, FL1, is further enhanced by placing the carrier frequency 20-dB down the filter slope as recommended by Collins.

The surplus crystal referred to may

not necessarily meet this requirement. Two or three could be tried, but ultimately the builder may wish to warp the crystal frequency, or he may wish to modify the crystal itself. I have found that regardless of the carrier position, greater roll-off of high frequency audio is desirable at the output of the device. Adding a simple series RC network (R = 10k pot, C = 1 μF) from pin 7 of the LM373 to ground allows more latitude in tonal balance.

I do not find that adding any of the suggested resistors really improves matters. There is an overall loss of system gain necessitating compensation with the transmitter gain control. Instability is not a factor in my device, but the suggested isolation may be helpful, especially at high frequencies. Concerning resistors, one which could be left out of the circuit diagram is incorrectly shown at the input to FL1 as going to ground. If retained, it should be connected to the other filter terminal.

The real problem in comparing the clipper to the compressor, it seems to me, has more to do with the limiting action of the transmitter alc than with any defects in the clipper. With or without the 2k ohm resistor suggested by Mr. Schreuer, severe clipping is possible (my data sheet indicates 70 ohms at pin 9). Heavy driving of the device will produce a solid carrier-like bar on an oscilloscope trace at the clipper diodes.

Power gain measured on a wattmeter appears to be about 4 dB with the compressor and 7 dB with the clipper when the input levels to the transmitter are equal. I note that Comdel claims 10 dB for their device, Magnum-Six, 6-dB, and DX Engineering, 5-dB power gain. But it is possible to drive the transmitter harder with the compressor and achieve 5 to 6 dB gain without flattopping. This observation is the basis of my conclusion, which has not changed.

With this condition I made the following table showing how the compressor and clipper treat pure vowel sounds and two voiceless consonants. English vowels are diphthongized and, therefore, unsustainable. So I have used the French vocalic system.

	compressor	clipper
a a	5.18 dB	6.02 dB
e e	3.97 dB	6.02 dB
i i	7.78 dB	10.00 dB
o o	2.55 dB	3.80 dB
u y	7.78 dB	9.03 dB
s s	11.76 dB	11.76 dB
t t	6.99 dB	6.99 dB
average	6.57 dB	7.66 dB

I have long felt, as does Mr. Schreuer, that the second filter is superfluous and could be replaced by a tuned circuit. The statement in the ARRL Radio Amateur's Handbook, *1970, page 258, ". . .a filter as good or better than the filter used to form the original ssb signal," in my opinion, only serves to confuse the issue.*

I used two filters because I had them. Two identical filters might actually have the disadvantage of reinforcing certain peaks and valleys in the passband.

Regardless of the refinements suggested, this device will add considerable authority to an otherwise unmodified signal.

Charles G. Bird, K6HTM

vertical antenna

Dear HR:

The article by VQ9N in the December, 1972, issue of *ham radio* on a single-element vertical antenna caught my eye immediately as it is similar to the antenna I have been using for about ten years.

When I moved to my present location I was unhappy with the antenna prospects. There is a 12,000-volt Edison line across the back of the lot about 60 feet from the shack with a maple tree in between. Putting up an antenna was a problem. After much head scratching, I finally took three sections of aluminum tv mast (about 30 feet), set it on a soft-drink bottle alongside the house with an insulator extended from under

the eaves of my two-story house. A single wire was run through a basement window about three feet to a tuner which is coax fed to the transmitter. I used the hot-water heating system for ground.

While reports are not spectacular, I get into the East Coast with good reports. My operating time is very limited so I have not calculated the engineering parameters of the system, but it is about as simple as you can get. I have used the antenna on 80, 40 and 6 meters but never on 20 where it would be most efficient. There are no guy wires. I would be the first to agree that it is makeshift, but it works. If you need an antenna in a hurry, or have a limited area, try it. My only problem is the tv service men who try to sell me a new tv antenna for the mast!

Paul R. Smith, W8FHB
Toledo, Ohio

Dear HR:

I was very interested to read the article by VQ9N on the use of a half-wave vertical antenna. For many years as VS4RS I used a 14-MHz version of the J-vertical described in the vhf section of the *ARRL Antenna Handbook.*

This antenna is essentially a half-wave vertical situated a quarter-wave above ground. My arrangement was a 50-foot guyed steel pipe with a 17-foot matching section at ground level, fed with 600-ohm open-wire feeder. No ground system is required.

Deducting two S-units for the rarity of the VS4 prefix, I am still convinced it was a very effective DX antenna.

Ron Shelton, 6Y5SR
Kingston, Jamaica

fetrons

Dear HR:

After reading your comments in the August, 1972, issue of *ham radio* regarding Fetrons, plus hearing some scuttlebutt from a ham in Seattle, I decided to give them a try.

I work for Pacific Northwest Bell in the Portland, Oregon, mobile telephone and two-way radio shop. We have some GE Progress line 150-MHz base-station receivers, and I thought we could upgrade them with the Fetrons. To make a long story short, Fetrons are very good attenuators at 150 MHz.

After going around in circles a few times I called Teledyne. Their first question was, "at what frequency are you using them?" When I indicated it was 150 MHz, they told me that was the problem. Contrary to the specification sheets, which call for a 500-MHz upper frequency limit, the upper usable frequency for current production Fetrons is 10 MHz.

The TS6AK5 and TS12AT7 Fetrons are being manufactured for telephone carrier systems, which have a top frequency of about 4.5 MHz. Teledyne indicated that Fetrons *could* be made to work at 150 MHz, but they do not at the present time.

As you pointed out in your editorial, most fetrons are designed for the lower frequencies. Before amateurs spend some of their hard-earned money on the current TS6AK5 and TS12AT7 Fetrons, they should realize that they are not usable much above 10 MHz.

Walter J. Loomis, K7BQE
Portland, Oregon

current limiting

Dear HR:

The December, 1972, issue of *ham radio* carried an article on adding current limiting to existing solid-state power supplies. I have used the circuit in the article, roughly as shown.

I tried other versions of current limiting and for simplicity's sake, I settled on one similar to the one in the article. As shown in the schematic, there is still one hangup that needs to be overcome. There is no protection against short circuits for the *current-limiting* transistor. Several deceased transistors attest to this fact. There are two causes for this. First, the instantaneous output voltage of the power supply appears across *R1,* the current-sensing resistor, at the onset of a short.

This places 15 volts, in this instance, across the base-emitter junction of Q1. This is several times the voltage rating of most transistors. To overcome this problem, placing a 1000-ohm resistor (ballpark figure) in series with the base lead to Q1 effectively gives it some current-limiting protection without unduly adversely affecting performance.

A second, and less likely to happen, transistor failure is due to the fact that Q1 is called upon to instantly discharge the electrolytic capacitor that is usually placed across the reference zener. This involves high peak currents with higher voltages and large value electrolytics. A 10- or 15-ohm resistor in series with the collector lead of Q1 limits this current. It is recommended that a fairly high beta transistor with at least 600 mW of dissipation be used at Q1.

The addition of these two resistors is a simple and inexpensive way to make the supply truly *current* limiting with very little adverse effect on operation.

Donald G. Cheshier, K5MKO
Garland, Texas

cooling fan error

Dear HR:

In the November, 1972, issue of *ham radio* the *ham notebook* section contains an erroneous concept of the nature of reactance. Author WB8IUF, in an attempt to lower the voltage supplied to a cooling fan, placed a capacitor in series with it. This is legitimate in itself, but he has fooled himself.

A capacitor is *not* equivalent to a resistance value equal to its reactance. The impedance also has a phase angle of 90 degrees. This means that the total impedance of the fan-capacitor combination is not the simple sum of the two impedances, but is the vector sum. This can be found from

$$Z_T^2 = R_F^2 = X_C^2 = 1200^2 + 442^2$$

$$Z_T = 1280 \text{ ohms}$$

where Z_T is the total impedance, R_F is the resistance of the fan and X_C is the reactance of the capacitor. In this case, where the total impedance is 1280 ohms, the current is 93.8 mA and the voltage across the fan, V_F, is 112.5 volts. Thus, V_F is considerably higher than that given by WB8IUF.

To find the correct value capacitor for this application use the following equations:

$$Z_T = \frac{V_T}{I} = \frac{120 \text{ V}}{75 \text{ mA}} = 1600 \text{ ohms}$$

$$X_C^2 = Z_T^2 - R_F^2 = 1600^2 - 1200^2$$

$$X_C = 1060 \text{ ohms}$$

At the line frequency of 60 Hz, 1060 ohms capacitive reactance is provided by a 2.5-μF capacitor. That gives the desired 90 volts across the fan.

Terry Conboy, WB6GRZ
Redwood City, California

yaesu spurious signals

Dear HR:

Reference, "Spurious Signals with the Yaesu," December, 1972, issue of *ham radio,* page 69. Unfortunately, the article by K6KA was incomplete with regard to which units of the FTdx560 can be affected by the spurious radiation problem.

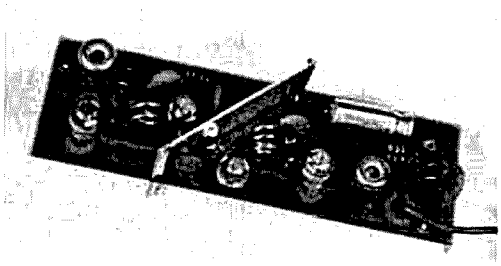
Starting with serial number 30001, the vfo frequency range in the FTdx560 was changed from 8400-8900 kHz to 8700-9200 kHz. The resulting change in local-oscillator frequencies produced a heterodyne with the second harmonic of the 3180 kHz i-f in some units to produce the spurious output. This was eliminated by addition of the 6358.6-kHz crystal.

All FTdx560s manufactured after introduction of the FTdx570, and all FTdx570s, have this circuit modification incorporated during production.

James Young
Spectronics
Signal Hill, California

new products

450-MHz mosfet preamp



Topeka FM Engineering has announced their new line of dual-gate mosfet preamplifiers. Their new model HF450 is designed for use in the range of 406 to 470 MHz, and is available in three models: 406 to 430 MHz, 430 to 450 MHz and 450 to 470 MHz. The rf voltage gain of these amplifiers is typically 15 dB with a 10-to 15-volt dc power supply. Noise figure is typically 4.5 dB. Superior cross-modulation performance and the greater dynamic range of the mosfet greatly reduces spurious responses. Each mosfet gate is protected by back-to-back diodes.

These new preamplifiers feature rf shielding of input and output circuits on both sides of the printed-circuit board and silver-plated G11 epoxy-glass boards for high performance. The preamplifier comes complete with all mounting hardware, rf jumper and detailed instructions. The HF450 is priced at $29.95 shipping prepaid. Also available is the model HF450 MO for Motrac radios. For more information, write to Topeka FM Communications and Electronics, 1313 East 18th Terrace, Topeka, Kansas 66607, or use *check-off* on page 110.

free electronics catalog

One of the biggest problems for the amateur home builder is obtaining the necessary electronic components for his projects. This problem is particularly bad if you don't live in or near a large metropolitan area. The new 1974 Olson Electronics catalog will solve many of these problems. This new, illustrated catalog, which features over 8000 quality-tested items, is free for the asking. In addition to a complete line of electronic components, transistors, ICs, etc., Olson carries the best of name brands in amateur radio equipment, antennas, test equipment, kits, electronic calculators and stereo and four-channel sound systems. For your free copy, write to Olson Electronics, Dept. HH, 260 S. Forge Street, Akron, Ohio 44327, or use *check-off* on page 110.

tube substitution handbook

Since 1960, there has been a tremendous increase in the number of American and foreign models of home-entertainment equipment. Consequently, an expansion of more than 600 percent has taken place in the number of available tube substitutions. This new and up-to-date edition of the Howard Sams *Tube Substitution Handbook* has kept pace with this rapid expansion by listing over 12000 direct replacements for all types of receiving tubes.

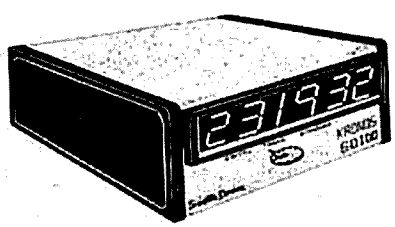
For convenience, this handy guide is divided into seven informative sections. Section 1 presents a cross-reference of all American receiving tubes, and section 2 lists picture tubes and their recommended

More Details? CHECK–OFF Page 110

substitutes. Section 3 contains a cross-reference of subminiature tubes, while the fourth section consists of industrial substitutes for receiving tubes. The fifth section is a substitute listing for communications and special-purpose tubes. The final two sections feature cross-references of American and foreign tubes.

There are easy-to-follow instructions accompanying each section that help you make proper tube substitutions and that explain how to cross-reference between sections for other substitutes. This guide fills the need of electronic experimenters, radio amateurs and service technicians who desire quick and accurate information for making suitable vacuum-tube substitutions. 96 pages, softbound. $1.75. A companion pocket-size volume which will fit into your pocket or your tube caddy is also available in a twin-pack which includes the regular sized volume, $2.25. Order from Comtec Books, Greenville, New Hampshire 03048.

digital chronometer kit



The Kronos KR100 chronometer kit features an LSI National clock chip, and a 32-page brochure with pictorials and easy-to-understand, step-by-step instructions. The chronometer includes 3 setting controls, 1-hour per second, 1-minute per second, and hold button. Easy-to-change from 12 to 24 hours, 4 to 6 digits, 50/60 Hz operation.

There are 3 models to choose from: 7-segment MAN-3 type LEDs, 6-digit kit, $47.00; 7-segment MAN-1 type LEDs, 6-digit kit, $69.95; and 7-segment Nixie type tube kit for $47.00. Available from Poly Paks, P.O. Box 942H, Lynnfield, Massachusetts 01940. For more information use *check-off* on page 110.

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100	.06	.08	.16
200	.08	.10	.20
400	.12	.14	.28
600	.14	.16	.32
800		.20	.40
1000		.24	.48

NEW

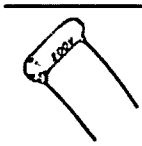
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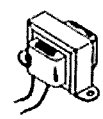
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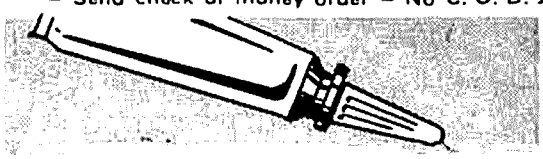
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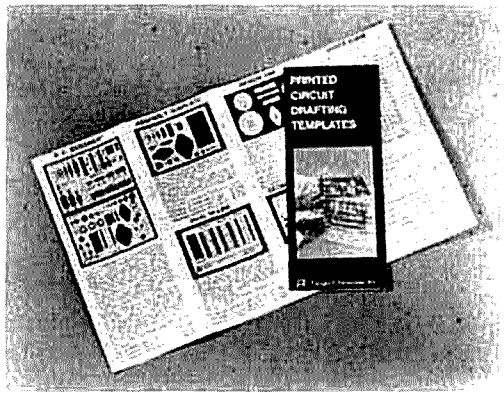
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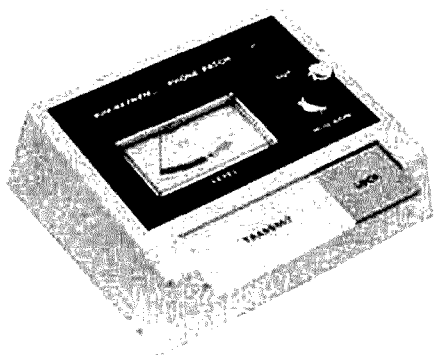
Oneida ELECTRONIC MFG, INC.
Meadville, Pa. 16335

drawing template catalog



Tangent Template, Inc. has published a short-form catalog that gives quick reference to a comprehensive line of user-designed drafting templates for printed-circuit design work. A total of fourteen different templates are described and illustrated, including templates for 1:1, 2:1 and 4:1 reproduction ratios. If you design PC boards, these templates should prove to be very useful. For a copy of the catalog, write to Tangent Template, Inc., Post Box 20704, San Diego, California 92120, or use *check-off* on page 110.

phone patch



New from Radio Shack is the Realistic Phone Patch at a price which the company says brings this useful accessory within the reach of any amateur's budget.

A phone patch provides an interconnection between your station's equipment and the telephone system, making it possible to place or receive telephone calls through a base station and relay them to another station which does not have access to a telephone. Phone patches

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have been of great use during civil emergencies, in providing communications in disaster areas, and often as a means for servicemen overseas to talk with relatives at home.

The Realistic Phone Patch is priced at $19.95 and comes complete with 15-foot telephone leads, three-foot transmitter lead and installation instructions. It features a built-in VU meter, gain control and locking push-to-talk bar. Not for use with transceivers employing electronic switching.

Realistic products are available at more than 1650 Radio Shack and Allied Radio Stores in all 50 states and Canada, and through Radio Shack Authorized Sales Centers, nationwide. For more information, use *check-off* on page 110.

rotary log-periodic antenna

Recently, KLM Electronics began manufacturing and marketing a line of antennas designed by Oliver Swan. This line includes a variety of high-performance vhf antennas that cover the spectrum from 50 to 520 MHz. The most recent addition to this line is a rotary log-periodic antenna that covers from 13 through 30 MHz. This antenna covers not only the 10-, 15- and 20-meter amateur bands, but MARS frequencies, the shortwave broadcast band and CB. Performance is equivalent to a three-element Yagi on any frequency between 13 and 30 MHz.

The KLM 13-30 log periodic uses seven elements and provides 9.2 dB over an isotropic. Front to back ratio is 12 dB. The input impedance is 50 ohms (a 4-kW balun is supplied), and maximum vswr at any point between 13 and 30 MHz is 2.0:1. The antenna weighs 76 pounds and has a boom length of 29.5 feet. The antenna is priced at $289.00, FOB San Jose, California. For more information, write to KLM Electronics, 1600 Decker Avenue, San Martin, California 95046, or use *check-off* on page 110.

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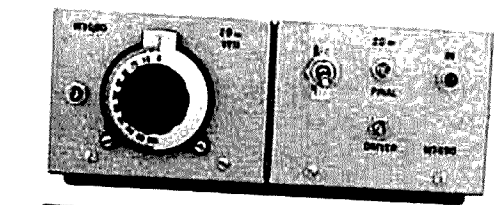
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ham radio magazine

NOVEMBER 1973



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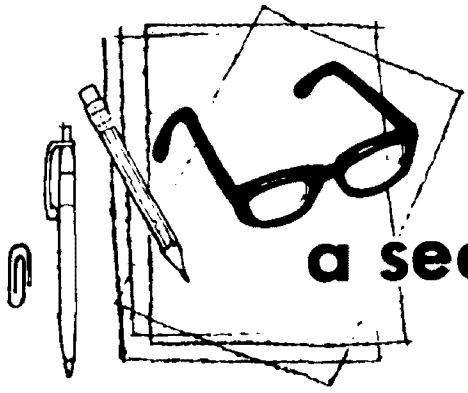
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a second look

by jim fisk

Because of an excessive number of problems being encountered by the FCC with amateur repeater license applications, and the lengthy processing delay, the Commission has extended the deadline date for all amateur stations licensed prior to October 17, 1972, which were automatically retransmitting radio signals from other amateur stations, *and* for which a timely and sufficient application has been filed. An application will be considered as being timely if it was received by the FCC on or before August 30, 1973.

According to the Commission, there apparently has been some confusion among amateurs as to the actual effective date of the rules adopted in Docket 18803. The FCC has reiterated that the rules became effective on October 17, 1972, and further, that full compliance was expected as soon as possible but not later than June 30, 1973. At the request of the American Radio Relay League this period was extended to August 30, 1973.

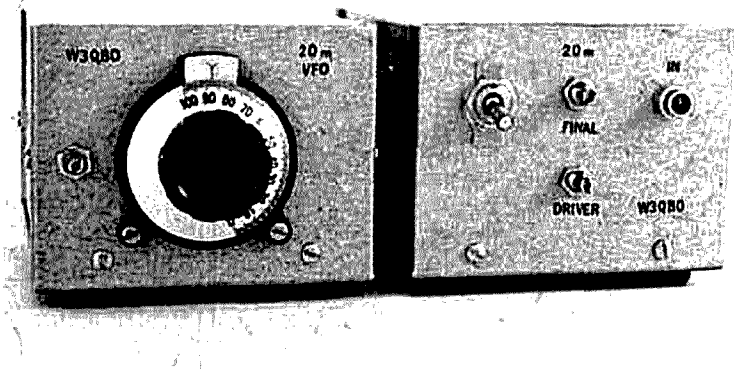
The FCC adheres to the view that all licensees have had adequate time in which to modify their repeater stations to fully comply with the rules set forth by Docket 18803, although there may not have been sufficient time to obtain the licensing authority for a repeater station, control station and/or auxiliary link station. Therefore, although the licensing deadline has been extended (apparently indefinitely), amateurs operating repeaters under previous authorization are cautioned that their operations must otherwise fully comply with the rules. The Commission has also pointed out that licensees and control operators of stations not operated in compliance with the rules of Docket 18803 are subject to appropriate enforcement action.

The FCC has complained that the main problems contributing to the lengthy processing delays are lack of standardization, failure to supply the required information and failure to present the information in a manner permitting rapid processing. In as much as there has been considerable confusion as to what information was actually required and the Commission was apparently reluctant to supply any guidelines, this is not surprising. After a considerable amount of arm twisting a few suggestions were forthcoming but they were nebulous at best. To add to the muddle, FCC staff members reached the epitomy of bureaucratic vacillation by approving one application and then disapproving another which used exactly the same format. If the problem wasn't so serious it would be amusing.

The Commission is now, finally, developing suggested application forms that are designed to eliminate the most frequently encountered errors. This should have originally been done at the time they adopted Docket 18803 — it would have saved everyone a lot of grief. Although it has not yet been decided if these suggested forms will be adopted as official FCC forms, properly prepared license applications using them will be acceptable for processing. You can also help by using universally accepted terms and standard symbols in your repeater license applications. If there is any question about an abbreviation, for example, spell it out — terms commonly used in one part of the country may mean something completely different somewhere else.

In the meantime, if your repeater was licensed prior to October 17, 1972, *and* you have filed a new license application prior to August 30, 1973, you can continue to operate your repeater under its present license *providing* it complies with all the rules of Part 97 of the regulations, as amended by Docket 18803.

Jim Fisk, W1DTY
editor



low-power solid-state vfo transmitter for 20 meters

Complete construction details for a two-watt QRP transmitter and Vackar vfo for CW operation on 20 meters

C. Edward Galbreath, W3QBO, 8326 Still Spring Court, Bethesda, Maryland

This article describes the construction of a 20-meter solid-state transmitter using a jfet Vackar vfo with a power output of two watts. This QRP transmitter has been used successfully for several months in many contacts, ranging geographically from New Zealand to the USSR. This has been done in a relatively few hours of operating time and — surprisingly — without the benefit of a beam; I use a 130-foot dipole, center-fed with open-wire line.

This past winter my kilowatt friends, skeptical as usual, challenged me to get on the air and try to work DX in a DX contest with this rig. I accepted their challenge, and despite lack of contest experience, I gave it a try for about two hours one afternoon during a recent DX contest. To the astonishment of all concerned I worked VP9HC, OK1TA, HB9KC, I5CFY, HA5KBM and I3ASE. Such accomplishments are only mentioned to show what can be done with two watts — I hope that more amateurs will give QRP a try.

The transmitter is built as two units, each in a 3x4x5-inch Minibox. The vfo, with the three buffer stages, is housed in a separate box only to meet my own

needs. The entire transmitter can easily be built in a single Minibox of appropriate size if you prefer.

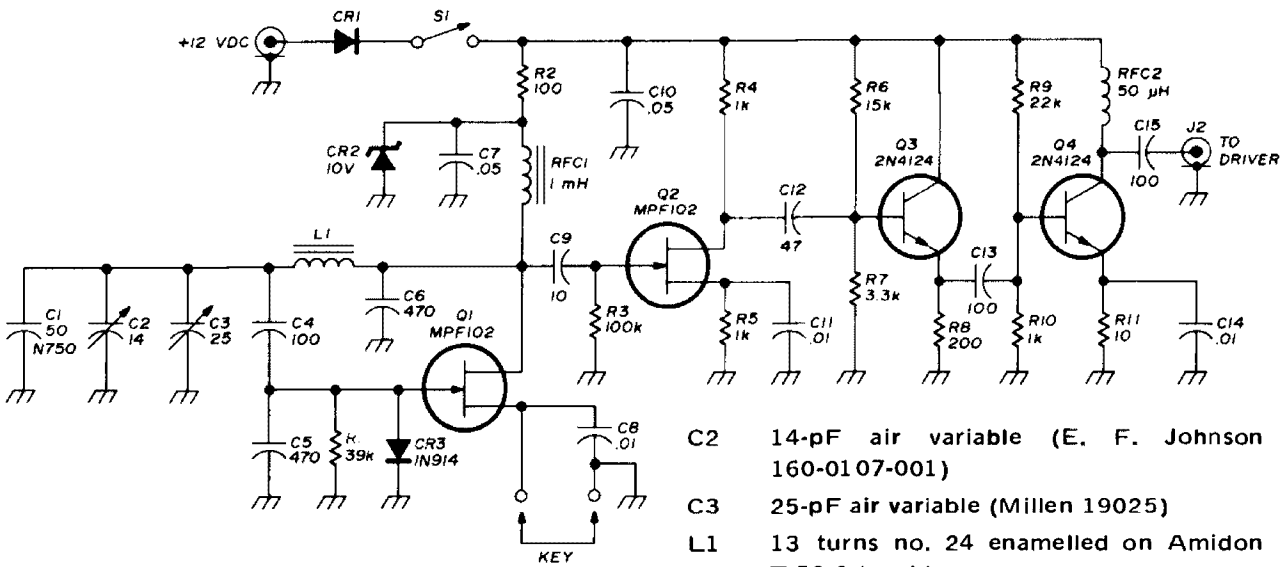
vfo

The schematic of the vfo is given in fig. 1. Except for the values of the oscillator's frequency-determining components and the addition of a buffer-amplifier stage, the circuit is the same as that used in my 80- and 40-meter versions.[1] The two buffer stages and the buffer-amplifier provide excellent isolation of the oscillator from the driver and final. There is only a slight frequency pull by the final amplifier even though the transmitter operates straight through.

tor, C1, is mounted on variable capacitor C3.

A clamping diode connected between the gate of Q1 and chassis ground has been added to the oscillator circuit. Wes Hayward, W7ZOI, when observing the gate voltage of an MPF102 on a high-frequency oscilloscope, found that, without the diode, the gate potential increased to the power supply potential on positive peaks.[2] The diode limits positive excursions to +0.7 volt, thus preventing conduction in the gate of the jfet and possible instability of the oscillator stage. Any small silicon diode such as a 1N914 will do the job.

All components of the oscillator stage



C1 50-pF, N750 temperature coefficient ceramic

C2 14-pF air variable (E. F. Johnson 160-0107-001)

C3 25-pF air variable (Millen 19025)

L1 13 turns no. 24 enamelled on Amidon T-50-2 toroid core

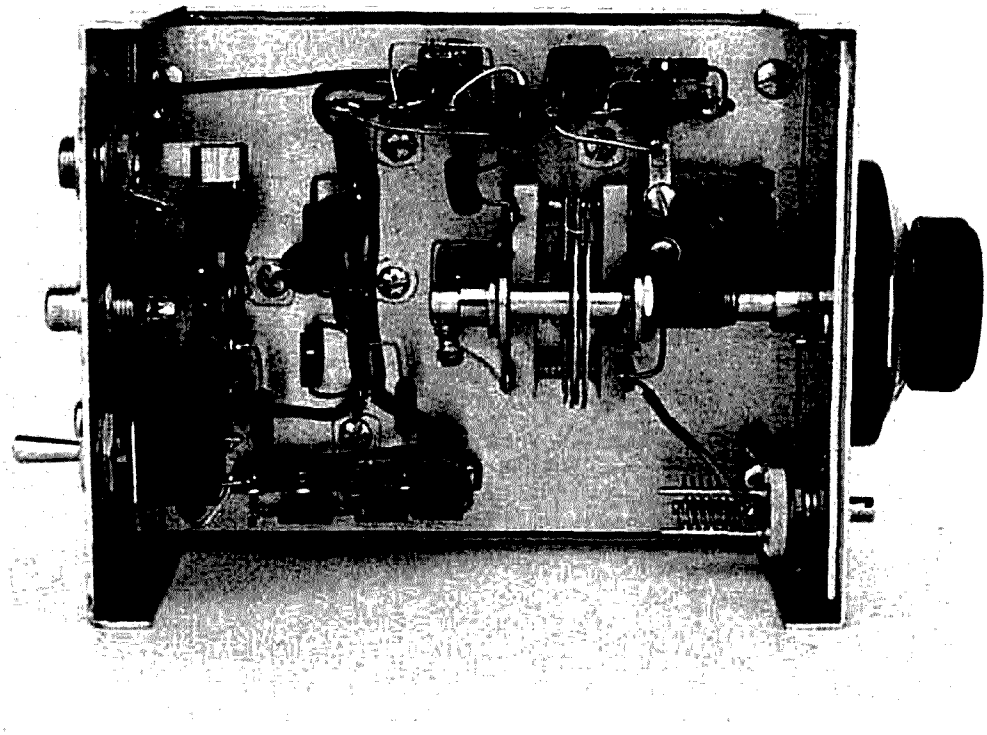
RFC1 1-mH rf choke (Millen J300-1000)

RFC2 50-µH rf choke (Millen 34300-50)

fig. 1. Circuit for the solid-state 20-meter vfo which tunes from 14.0 to 14.2 MHz is based on the very stable Vackar design. Protective diode CR1 is any silicon rectifier.

Frequency pull is so slight that there is no need to compensate for it when zero-beating a signal with the driver and final turned off.

The feedback inductor, L1, consists of 13 turns of number-24 enamelled wire, wound tight and evenly spaced on an Amidon T-50-2 toroid core. To assure short leads, the inductor is mounted on a perforated board on a standoff insulator between the tuning capacitor, C3, and the front panel of the Minibox. The leads are connected to soldering lugs. The negative temperature coefficient ceramic capacitor, except the inductor, variable capacitors and the negative temperature coefficient ceramic capacitor, C1, are mounted on two terminal strips: a 5-tie-point strip and a 3-tie-point strip. The center tie point is grounded in each case. The two terminal strips are mounted in line as a single 8-tie-point unit. This represents a constructional change from the earlier 80- and 40-meter versions and permits easy assembly of the components on the two strips as a unit before attaching them to the chassis. One simple way to do this is to bolt the terminal strips to a narrow



Layout of the vfo and buffer stages. Inductor L1 is mounted on the perf board just above the main tuning shaft. Transistors and other components are mounted on terminal strips. Key jack, on-off switch and output *connector are mounted on rear panel, to the* left.

piece of aluminum just as they would be mounted on the chassis. This unit can then be held in a vice while you attach the components.

To strengthen the Minibox, an aluminum chassis is built into it. The ends of the chassis are cut from half inch right angle aluminum stock available in most hardware stores. The aluminum angle strips are attached to the front and back panels at the base of the box and an aluminum plate is bolted to them. This chassis can be removed from the Minibox after fitting for easy assembly of the terminal strips, the tuning capacitor and the inductor.

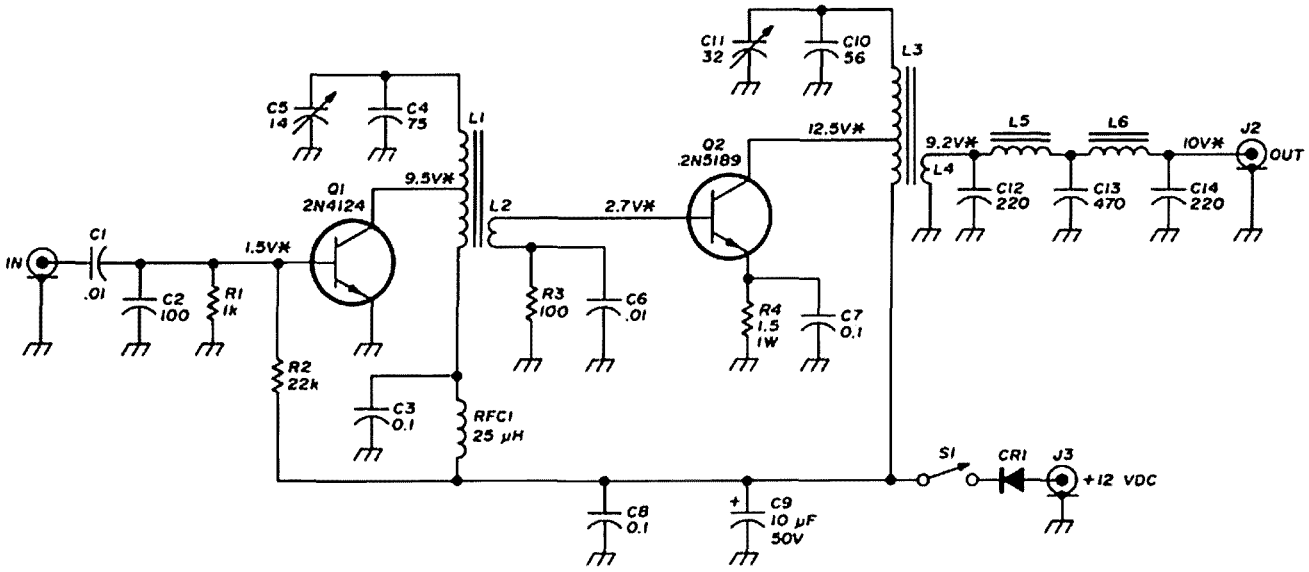
A piece of aluminum is used to reinforce the *front* panel. It is held in position in the box against the front panel by the machine screws that fasten the chassis in place and by the dial and the trimmer capacitor, C2. Be sure that this piece, as well as the chassis, are cut just small enough to permit the cover to slip on the box. The extra reinforcement may be omitted, however, if the additional strength and rigidity are not desired.

A two-inch vernier dial is used to adjust the main tuning capacitor. It is a common item found in many radio stores but can also be ordered by mail. Holes for the dial can be marked for drilling only after the chassis has been fitted and the main tuning capacitor is in place. A key jack, two phono jacks and an on-off switch are mounted on the rear panel. One phono jack is for battery or power supply leads, the other for rf output.

driver and final

A Motorola 2N4124 transistor is used to drive an RCA 2N5189 in the final amplifier to 2 watts output (see fig. 2). Both transistors are inexpensive, priced in the 70 to 80 cent range. The 2N5189 is a very efficient transistor but it seems to have been overlooked by amateurs.

The 2N4124 driver operates as a class-B amplifier with its base biased to approximately collector-current cutoff. With no signal applied to the base, col-



C5 14-pF air variable (E. F. Johnson 160-0107-001)

C11 32-pF air variable (E. F. Johnson 160-0130-001)

L1 16 turns no. 24 enamelled on Amidon T-50-2 toroid core, tapped 6 turns from B+ end

L2 2 turns small insulated wire wound over B+ end of L1

L3 16 turns no. 20 enamelled on Amidon T-68-2 toroid core, tapped 3 turns from B+ end

L4 3 turns small insulated wire wound over B+ end of L3

L5 11 turns no. 20 enamelled on Amidon

L6 T-50-2 core

RFC1 25-µH rf choke (Millen J300-25)

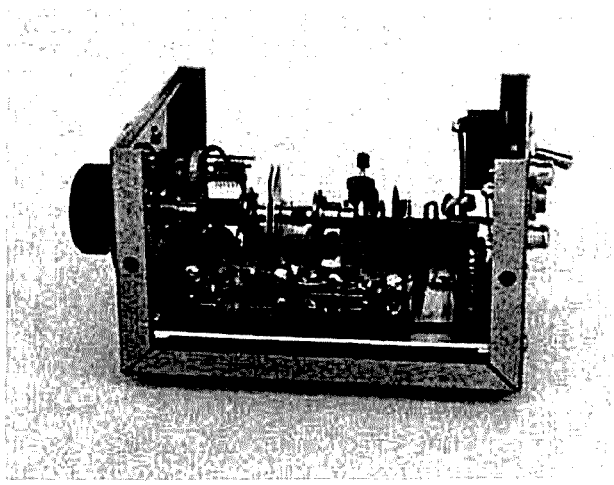
fig. 2. Solid-state driver and final provides 2 watts output on twenty meters. The voltages marked with an asterisk are the rms rf voltages at those points. Protective diode CR1 is any silicon rectifier.

lector current is near zero, thus minimizing current drain during key-up condition. As a class-B amplifier, the stage is easier to drive than if operated class-C. A capacitor divider consisting of C1 and C2 provides the correct impedance match between the vfo and the base of the 2N4124.

A fixed-tuned tank circuit is used in the driver stage. Once the tank is correctly adjusted for maximum output in the center of the CW portion of the band, no further adjustment is required. The primary, L1, of the tank coil is tapped 6 turns from the B+ end to assure a proper low-impedance match for the collector of the 2N4124. Two capacitors are used to tune L1, a 75-pF dipped mica and a miniature variable, C5, which is mounted on the front panel.

The secondary of the driver tank, L2, consists of two turns of small insulated wire wound over L1 at the B+ end. This provides a low-impedance match to the base of the 2N5189. A 100-ohm resistor in series with L2 and chassis ground helps protect the 2N5189 from damage.

The tank circuit of the 2N5189 final is similar to that of the driver stage. The collector is tapped to L3 at 3 turns from the B+ end. The variable capacitor, C11, is mounted on the front panel directly above C5. A double-pi network consisting



The vfo and buffer stages are built into a 3x4x5-inch Minibox. Extra chassis stiffness is provided by the 1/16-inch aluminum subchassis and panel stiffeners.

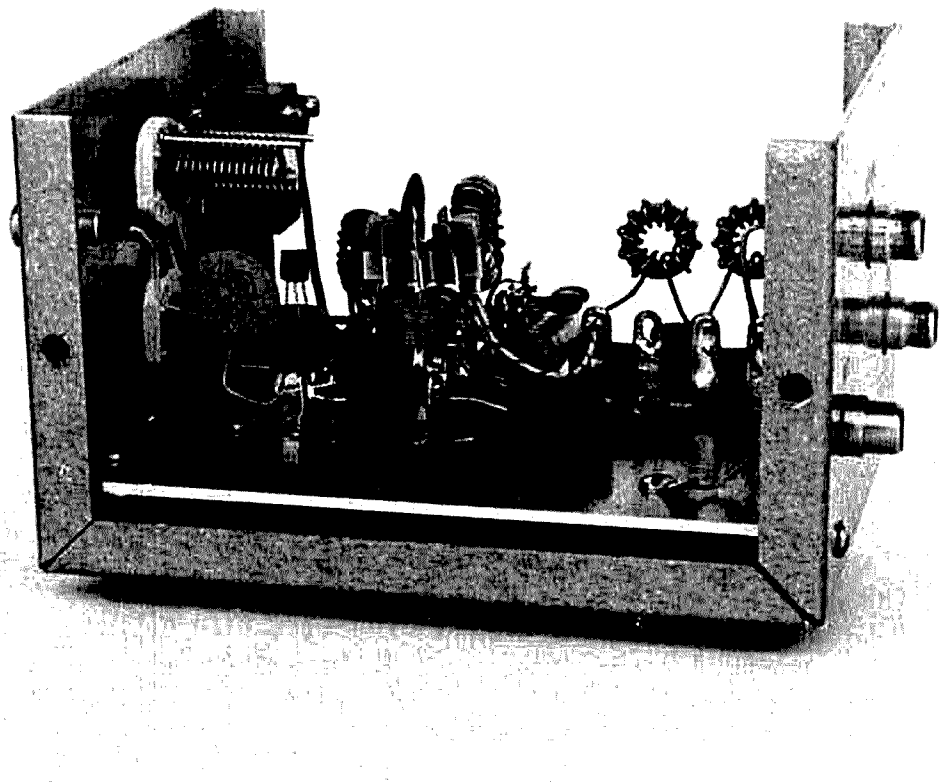
of L5, L6, C12, C13 and C14 is used in the output to assure good harmonic attenuation.

construction

Construction of the driver-final unit is similar to that of the vfo. An aluminum chassis is cut and fitted inside the Minibox using 1/2-inch right-angle stock and a piece of flat aluminum cut slightly smaller than the bottom of the box to permit the cover to slide in place. The chassis is then removed and all the components except the variable capacitors are mounted on it out of the box.

Components of each stage are mounted on two terminal strips — one 5- and one 3-tie-point strip — arranged as an in-line unit. The driver stage is near the front. The double-pi network is mounted at the back on a 4-tie-point terminal strip. This arrangement permits orderly assembly of components and short leads.

The Minibox should now be drilled for the on-off switch, input jack and the two miniature variable capacitors on the front panel and the output jack and the B+ jack on the back. You may also want to include an antenna jack for connection to the receiver if one antenna is used for both transmitting and receiving. This jack should be connected to the transmitter output through a 68-pF capacitor, with reversed silicon diodes to ground on the receiver side of the capacitor to protect the receiver against damage.



Like the vfo, the driver and 2-watt output stage are built into a 3x4x5-inch Minibox. Transistor Q2, a 2N5189, is provided with a clip-on heat sink.

testing

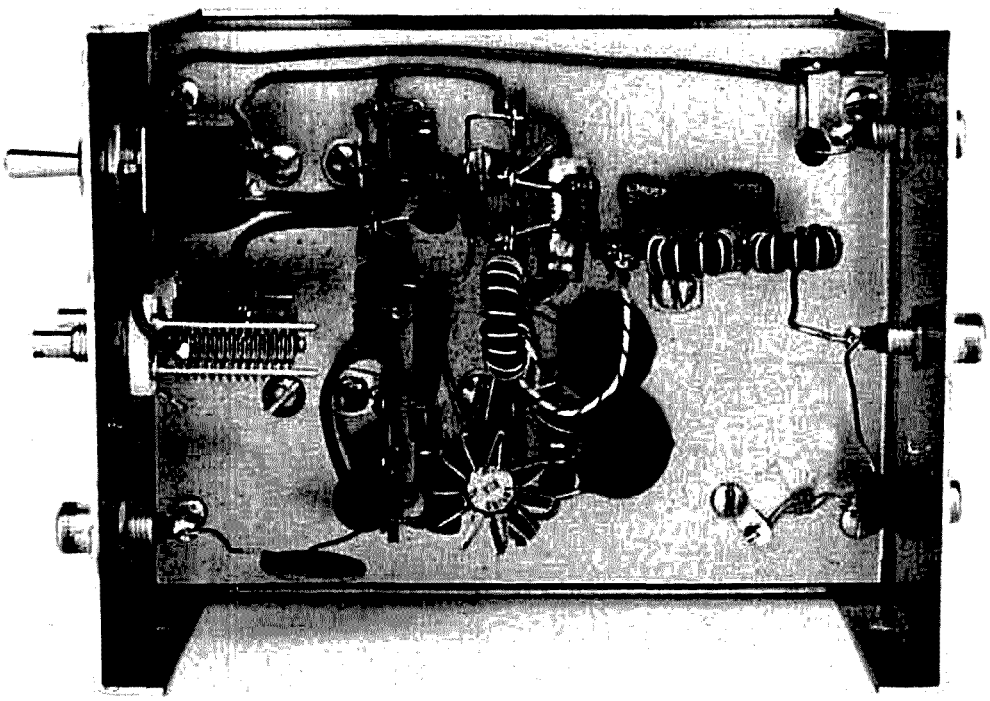
Check both units for possible shorts to the chassis before connecting them to a power supply (and before inserting the transistors, if you use sockets). If the units pass the short test — B+ to chassis — then insert the transistors and connect the output of the vfo to the input of the driver-final using a short, shielded jumper

cable. Output of the final should be attached to a 50-ohm dummy load; a two-watt resistor works well.

The units require 12 volts dc. I use a separate 12-volt lantern battery for each unit so as to contribute toward good voltage regulation for the vfo.

with your multimeter. If you have a sensitive reflected power and swr bridge meter, check for output. Adjust the front panel trimmer capacitors for maximum power output at the midpoint of the CW portion of the band.

This completes the adjustments and



Layout of the driver and power output stages. All components are mounted on terminal strips. Tank circuits L1/L2 and L3/L4 are mounted just to the right of the air variable. Inductors L5 and L6 are mounted on another terminal trip near the rear panel.

Now, turn on the power supply switch for the vfo and, while keying, listen for the signal in your receiver. When you have found the signal, set the dial for full capacitance of the tuning capacitor and use the air trimmer to set the low band-edge at 14 MHz.

Next, turn on the driver-final unit and again key the vfo. You should now hear a stronger signal. If you have an rf probe and a vtvm, you can check for rf voltages at points given in the schematic. If you do not have an rf probe, rectify the rf output with a germanium diode and capacitor and read the rectified voltage

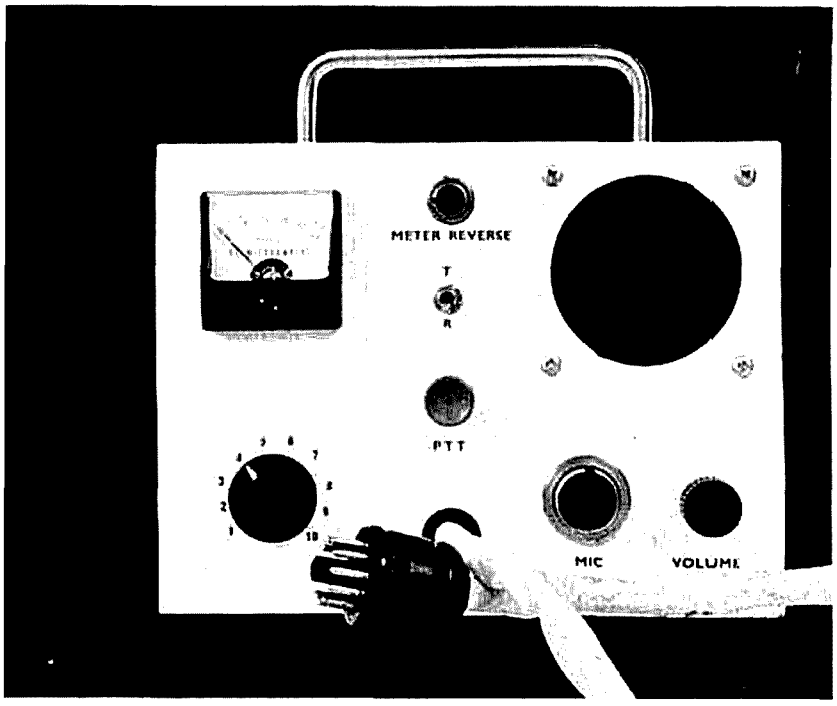
checking. The transmitter is ready to be put on the air. An antenna tuner is strongly recommended for matching the transmitter to the transmission line and for additional attenuation of harmonics. The extra resonant circuit provided by the antenna tuner also helps in receiving.

references

1. C. E. Galbreath, W3QBO, "A VFO for Solid-State Transmitters," *ham radio*, August, 1970, page 36.

2. Wes Hayward, W7ZOI, "A Second Generation MOSFET Receiver," *QST*, December, 1970, page 12.

ham radio



test set

for Motorola radios

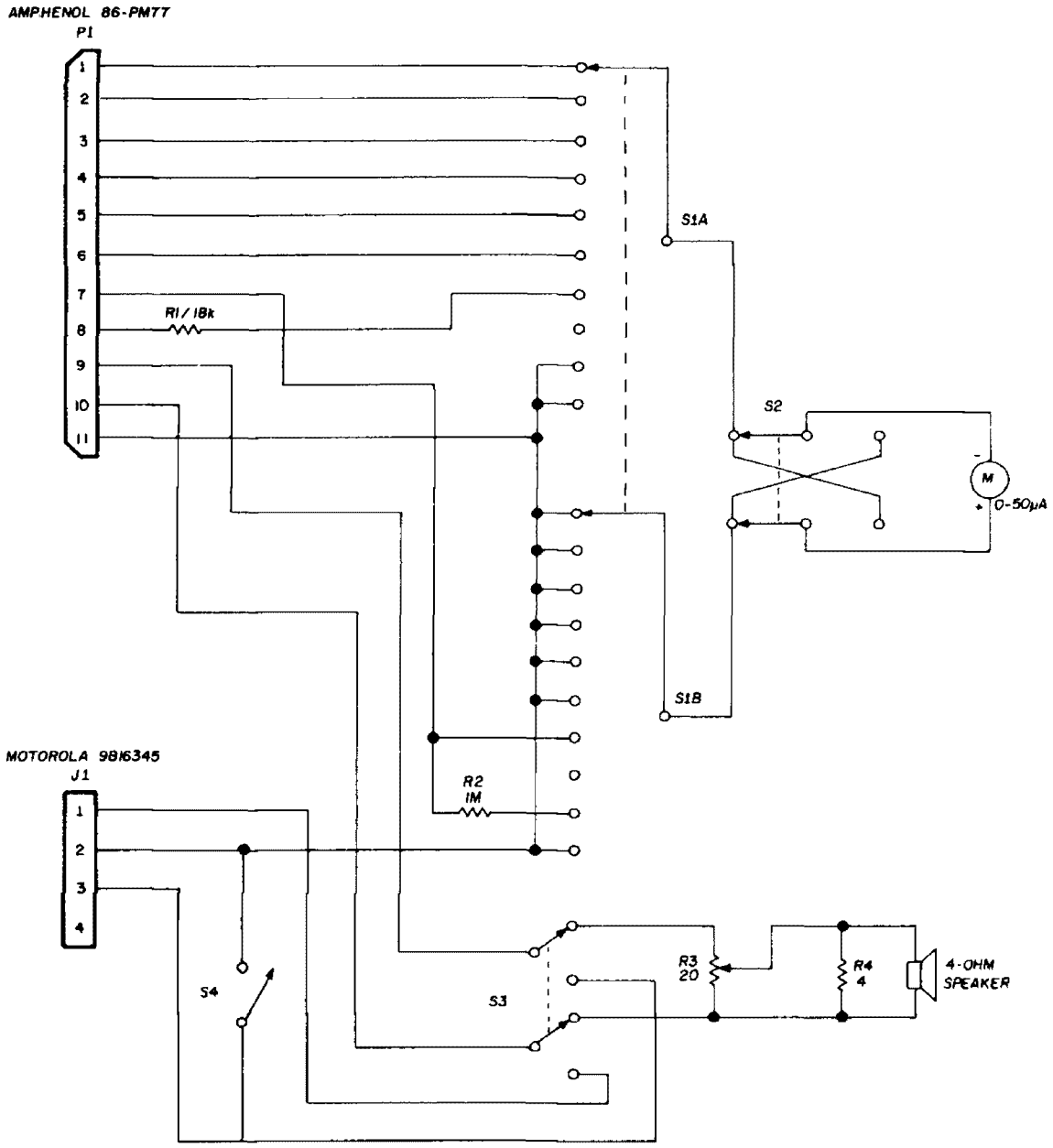
Complete construction details for a test set for most Motorola business-band fm radios

David L. Marshall, KØBKD, Albuquerque, New Mexico

Today is the day of vhf-fm communications. Although there are many amateur-band fm transceivers in use, retired business-band fm sets still hold a sizable portion of the market. This is due to their increased reliability resulting from their ruggedized construction suited to a mobile environment. Another reason is their reduced cost.

One minor disadvantage with old business-band radios is that they usually require specialized test equipment to keep them operating. The average amateur has no way to maintain his own equipment and has to rely on a friend in the two-way radio business. This is fine until the radio quits and test equipment is not available for one reason or another. In my opinion, half the fun of ham radio is maintaining your own equipment.

In this article I will describe a simple but comprehensive test-set for Motorola fm radios. Although you can buy a readymade test-set from Motorola, they cost about $250, which is a bit expensive



M1	50 μA meter (Simpson 1212)	S1	2 pole, 10-position, non-shorting wafer switch (Centralab PA-1005)
R1	18k, ½ watt, 5%	S2	snap-action, momentary contact dpdt switch (Alco MSPE-206R)
R2	1 megohm, ½ watt, 5%	S3	miniature dpdt toggle switch (Alco MST-205N)
R3	20-ohm potentiometer, 5 watt	S4	spst momentary contact switch
R4	4 ohms, 2 watt		

fig. 1. Circuit for the test set for Motorola vhf-fm radios. Adapter cable for Motrac sets is shown in fig. 2.

for a piece of equipment that isn't (hopefully) used too often.

Fortunately, many years ago Motorola showed great insight by designing their radios to be compatible with a universal test-set. Today, the same test-set will align 80Ds, 140BYs, Twin-Vs, T-Powers, utility base stations, Motracs, Motrans, Mocoms and Micors. There are many different models included under these different generic names so it is obvious that a test-set is a most handy piece of

equipment to have, regardless of what Motorola radio you may have. The few exceptions are the H23XXX hand-carry, the HT-200, the HT-220, the PT-200, PT-300, etc. Due to space limitations, these sets do not have meter sockets and are aligned with a volt-ohm-meter.

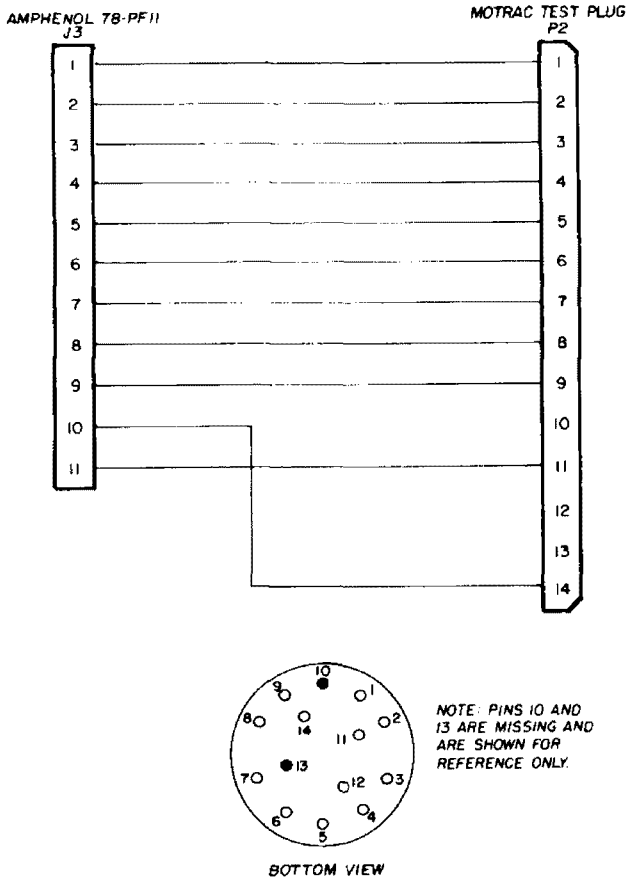


fig. 2. Motrac adapter cable for use with the test set shown in fig. 1. In the bottom view of the Motrac test plug, solid dots are pins, open dots are holes.

test-set

The basic test instrument for checking Motorola radios is a 50-μA meter. By switching in suitable multipliers, a voltmeter can be built which will monitor all test points. There are two additions to this basic meter which make it even more valuable. One is bringing the receiver audio out to a speaker built into the test-set. This facilitates working on trunk-mounted radios where the normal speaker is some distance away. In the commercial version there is also an audio voltmeter used for making signal-to-noise measurements. For reasons of simplicity this feature was not included in the unit described here. Another useful addition to the basic test-set is the provision for transmitter push-to-talk and a microphone connector. These items are extremely useful when you are working on a transmitter with a remotely located microphone.

Although there is nothing critical about the construction of the test-set, there are a few considerations which are worthy of mention. From experience I have found that a cable about three feet long is optimum. To align Motrac, Motran, Mocom and Micor radios an adapter is necessary. This converts the normal eleven-pin test-set plug to a plug similar to a nine-pin tube except that it has four pins in the center. EECO logic modules have this kind of plug, or the adapter can be purchased from Motorola (TKN6025A, price $9.50). If you do not anticipate working on any of these radios this adapter is not needed as the earlier radios used an eleven-pin socket which is compatible with the test-set plug.

Another area of flexibility is in the meter reversing switch. It can be incorporated into the selector switch so that position +4 is normal and position -4 is reverse. Since the reversing switch is used only for discriminator alignment it is recommended that the meter reversing switch be a separate pushbutton switch so that it is not necessary to switch through the reverse position every time. It is also recommended, for reliable operation, that the switch be a good quality, snap action microswitch. Quality switches are a must since changes in contact resistence, encountered with inexpensive switches, can change meter readings.

The selector switch should be a non-shorting type, also of good quality. I happened to use a shorting-type switch because I had it on hand; however, when using a shorting-type switch, if the unit is switched while measuring PA plate current, the power supply fuse will blow because of the momentary short circuit.

The fm test-set will be only as good as the meter, so a good quality meter movement is a must. I recommend using a Simpson meter since Motorola uses this

type in their test-set so the difference in internal resistances should be small. Also, the largest available meter should be used to get the maximum needle swing – some of the indications are quite small.

The microphone connector is a special connector similar to an Amphenol 91-PC4F. Occasionally these can be found on a junked Motorola control head. If you can't locate an old one, connectors can be ordered from Motorola (9B16345, price $.95).

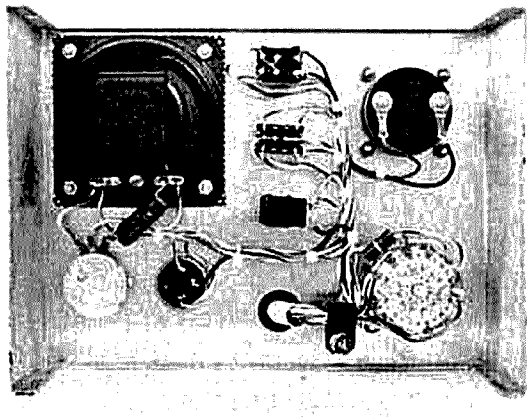
I built my test-set into a Bud CU-3009A Minibox, 3½ x 6 x 8 inches. This size is large enough to allow for modifications such as a larger meter, yet is small enough to be easily portable. The unit has seen a great deal of service in mobile maintenance and working on repeaters. It has always been easily portable.

motorola alignment

For those of you who are not fortunate enough to own a manual on your particular radio, receiver alignment is very straight-forward. However, because of the different number of multiplier stages used in the various frequency ranges, Motorola has chosen to base the alignment on the power amplifier stage. The PA plate is always position seven. The PA drive is position six, the last multiplier is position five, etc. In a low-band radio the switch will go back only to position four. A uhf radio will go back to position one.

Receiver alignment is started by aligning the discriminator for a zero output with an on-frequency signal. There are several methods in use, but generally the primary is peaked with the secondary detuned and then the secondary is tuned for a zero voltage output. It is best to use a 455-kHz signal rather than relying on a received signal. When this is done the 455-kHz i-f stages are aligned again using a 455-kHz signal. The signal level should be kept below the point of saturation on meter position one. If the receiver is badly detuned, position two may be used for a more sensitive indication.

The 12-MHz i-f is aligned by coupling a signal at 12 MHz to the first mixer. The high i-f is then aligned again using position one. At this point an on-channel signal should be copiable, even though weakly. Using this on-channel signal, the rf amplifier should be aligned. When this is done the signal should be further reduced and, using position two, the complete alignment should be touched up for the best sensitivity.

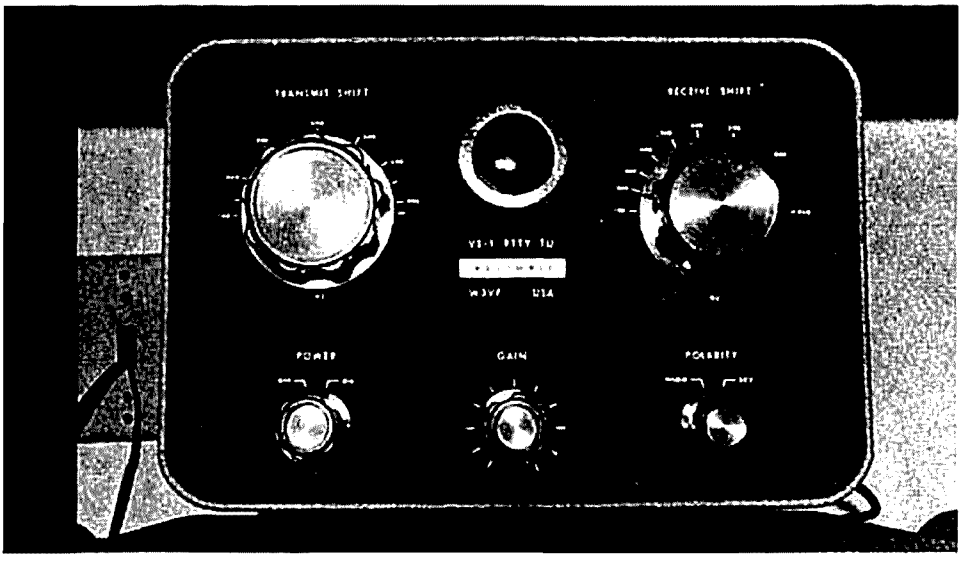


All components for the Motorola test set are mounted behind the front panel.

The final step in the alignment procedure is to adjust the received frequency. This is done by adjusting the first oscillator trimmer for a zero-voltage indication at position four.

Transmitter alignment is simply a matter of tuning each stage for maximum meter indication with the exception of position seven. Position seven is adjusted for a dip in the PA plate current as in any class-C stage. If the loading and coupling controls are adjusted for maximum and then backed off to about 80% the output power will be reasonably close to what Motorola recommends for that particular radio. The final step is to adjust the transmit frequency. I use a 30-MHz frequency counter coupled to the output of the phase modulator. It is important not to load the oscillator as this will change the frequency when the counter is removed. By using a one-second time-base on the counter, you will obtain accuracy that is entirely adequate for amateur vhf operation.

ham radio



variable-shift RTTY terminal unit

And now the VS-1 — with continuously variable filter tuning for improved response in noise and interference

Keith Sueker, W3VF, 110 Garlow Drive, Pittsburgh, Pennsylvania

Phase 1 of my RTTY experience began with the gift of a Model 26 Teletype machine from Mac McKinley, W30B. During my first attempt to print something off the air, a simple single-tone detector and vacuum-tube keyer were used.[1] This system worked but with poor results in noise and interference. At this point I began to learn something about TU problems and their solutions.

Phase 2 was introduced by a version of Irv Hoff's ST-5,[5] which was modified to use tone pairs of 1000/1170 Hz and 1000/1850 Hz for narrow and wide shift respectively. These frequencies were chosen to allow the sharp CW selectivity in my SB-300 to be used for narrow shift without cluttering up the receiver with an odd-frequency BFO crystal for RTTY. This system unit worked well enough to allow some 20 meter DX.

Phase 3 began with some serious considerations of how to build a better TU in terms of noise and interference response. Advanced design units, such as the ST-6[6] with switched discriminator filters, do an excellent job on narrow shift but have an excessively broad response for marginal

receiving conditions on wide shift. The phase-locked loop units[7] have the same problem, since they are subject to interference through the whole lockup range of the vco. A solution is to use sharply tuned filters, in or ahead of the discriminator, which will pass only the individual tones of interest and their sidebands. When fixed-tuned filters are used, however, one runs into the problem of various amateur shifts and the many commercial stations that use shifts other than 170 and 850 Hz.

The ST-6 discriminator filters at 170-Hz shift are about as narrow as one might wish to use. The 850-Hz shift filters, however, are broad enough to provide a linear discriminator characteristic over 850 Hz. If we sharpen these filters, we can reduce the interference bandwidth, but we still must cope with the various intermediate shifts in use. Why not make one of the filters with a continuously adjustable center frequency? Then we can tune any shift.

the VS-1 TU

The filters are the key to the whole project. If one filter were continuously tunable, what kind of filter and tuning system could be used? Active filters using RC networks are popular, but high Q requires critical feedback adjustment and stability is questionable with simple techniques. Toroidal LC filters using 88-mH inductors provide excellent Q, but how does one tune such a device? To accommodate the shift range between 170 and 850 Hz, one filter must be varied nearly 700 Hz — a task clearly impossible for an ordinary variable capacitor. Also, resistive loading would have to be continuously variable to keep the noise bandwidth matched to the fixed filter.

The solution to these and other problems turned out to be relatively simple: both audio tones were increased to higher frequencies so that one of the filters could be tuned with a variable capacitor. Some calculations showed that a frequency of about 11 kHz would allow a sufficient tuning range with a 365-pF capacitor. At 11 kHz, the tuning range would be less than ten percent of center frequency, providing good noise balance without compensation.

By going to a high frequency, the absolute bandwidth would be increased by some four times over that obtainable

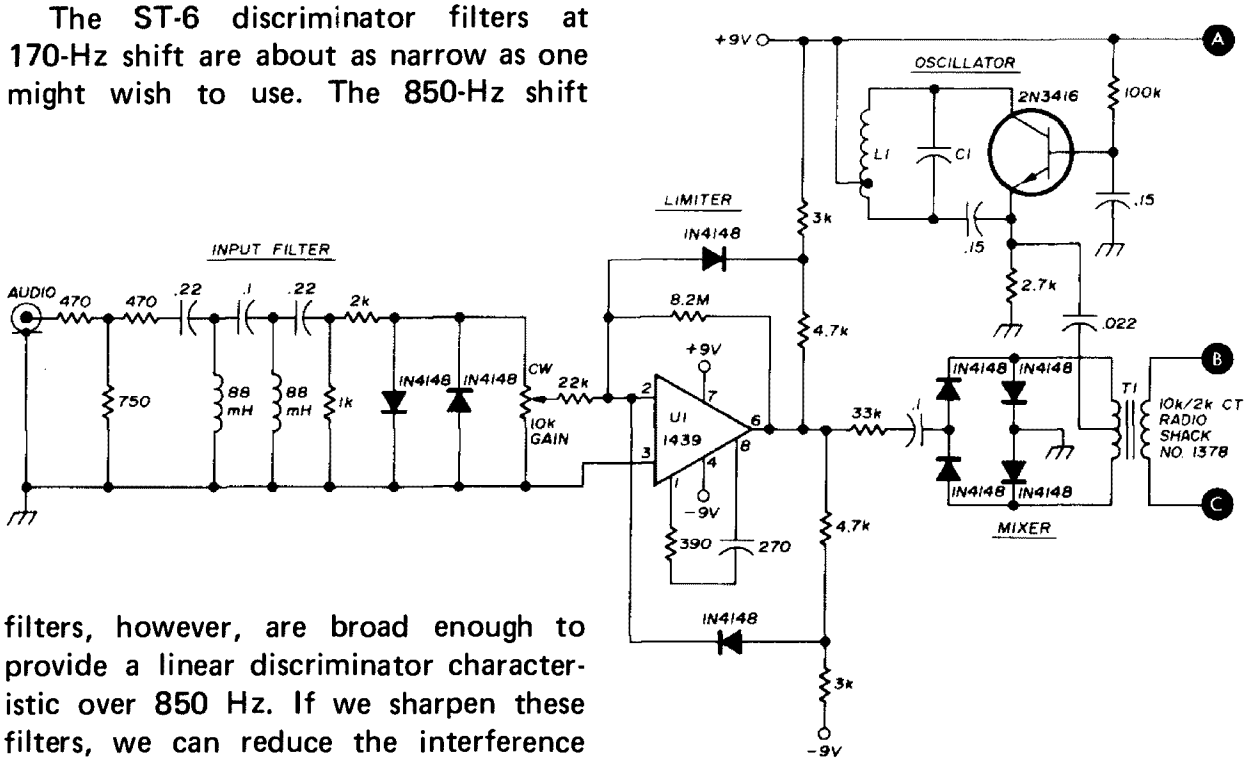


fig. 1. Input circuit and mixer for the RTTY terminal unit. For values of L1 and C1, see text.

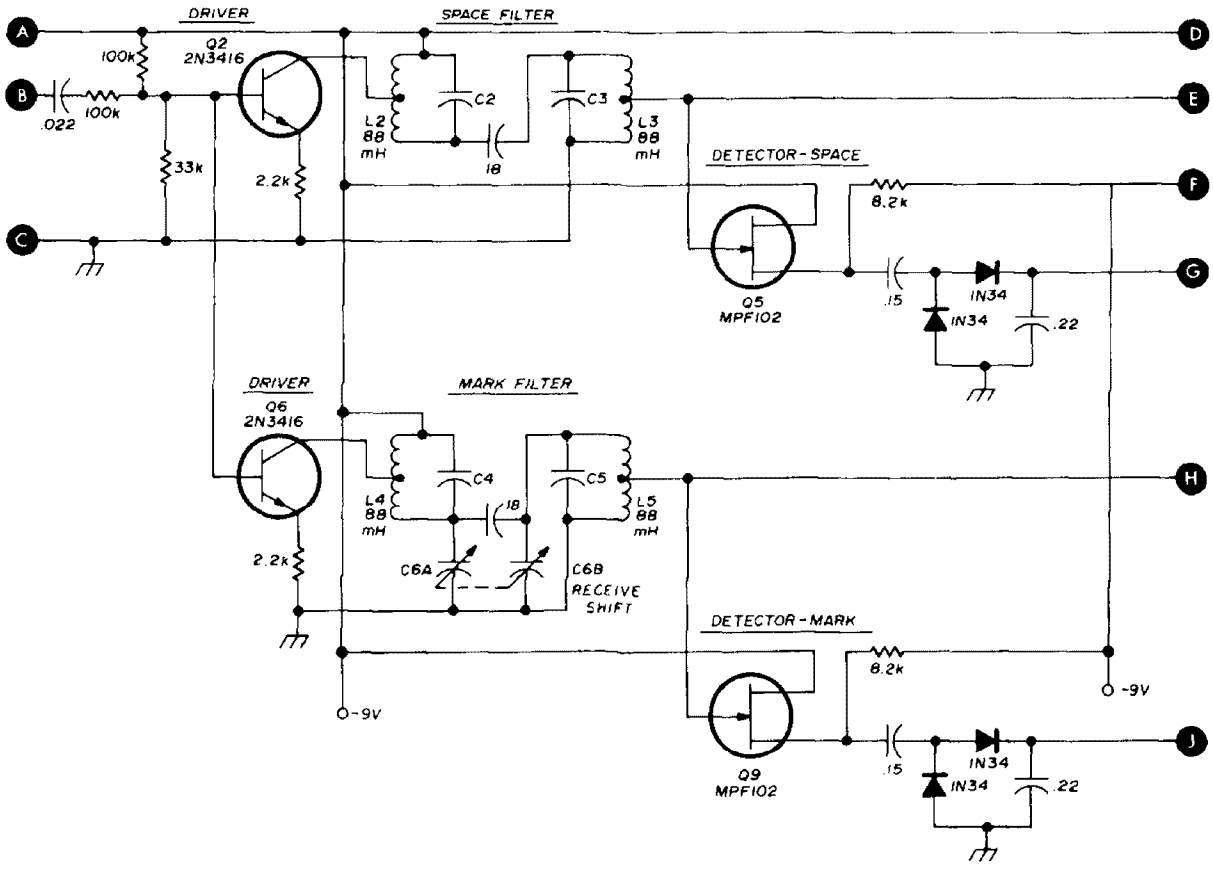


fig. 2. Mark - space filters and detectors for the RTTY terminal unit. For value of tuning capacitors, see text.

at the usual audio tones. This problem was resolved by using two filter sections with loose coupling and a dual-section capacitor for tuning. The selectivity turned out to be ideal.

input circuit and mixer

Whether using fm or limiterless a-m detection, it is important to restrict incoming audio to the frequencies of interest. Most modern receivers have i-f filters with sufficient high-frequency cutoff. However, the low-frequency response should extend no lower than the bottom of the i-f passband. Hence, a high-pass filter is used on the VS-1 input to reduce hum and noise (fig. 1). Toroidal 88-mH inductors are used, and the filter is designed for a low-frequency cutoff of 1000 Hz. The input pad allows satisfactory operation from any receiver output impedance. Measured attenuation is 1 dB at 1000 Hz, 13 dB at 700 Hz, 29 dB at 500 Hz, 38 dB at 300 Hz, and more than 40 dB for lower frequencies.

Op-amp U1 and the associated circuit form a high-gain stage, which can be operated as a linear amplifier or a symmetrical clipper (limiter) by varying the input gain control. The gain control could be replaced by a switch as in the ST-6; but a control provides more audio range.

Once the audio tones have been amplified and/or clipped, they are heterodyned to the filter frequencies by the local oscillator and balanced mixer. The oscillator is a Hartley circuit with a tap on L1, which is made by unwinding 50 turns from an 88-mH toroid, attaching a lead, and rewinding the turns in the same direction. Capacitor C1 tunes the oscillator to the desired injection frequency. (More about this later.) Use of an 88-mH toroid ensures that the oscillator and filters will track over a wide temperature range. The balanced modulator is conventional. It uses unmatched diodes and is coupled to succeeding stages through an inexpensive transistor interstage transformer.

filters

Transistors Q2 and Q6 are filter drivers (fig. 2). The high collector output impedance, coupled with feed at the center tap of the toroid, results in negligible filter loading. Each filter consists of two LC sections using 88-mH toroids with loose coupling provided by the 18-pF capacitors. Space filters L2,C2 and L3,C3 are fixed-tuned to the heterodyned space frequency. Mark filters L4,C4 and L5,C5 are identical to the space filters except for the dual variable tuning capacitor, C6, which tunes the filters to the heterodyned mark frequencies.

frequency selection

Up to this point discussion of the frequencies has been rather vague. This TU is basically a superhet receiver with dual i-f channels, one of which can be tuned. A nominal i-f must be used that can be varied by the frequency difference between maximum shift (usually 850 Hz) and minimum shift (usually 170 Hz) with a ganged capacitor of reasonable size. For a shift difference of 680 Hz and a dual 365-pF broadcast capacitor, the minimum possible filter frequencies are about 11,000 Hz and 11,680 Hz for the tunable filter. If the other filter is set at either 10,830 Hz (11,000-170 Hz) or 11,850 Hz (11,680 + 170 Hz), the entire range of shifts from 170 Hz to 850 Hz can be tuned with the variable capacitor.

space and mark. Common practice in amateur RTTY is to shift frequency down for space. If the receiver sharp filter is to be used, the lower of the two audio frequencies must fall at the lower end of the sharp passband – about 950 Hz in the Heath SB series. Narrow shift then has space at 950 Hz and mark at 1120 Hz, while wide shift has space at 950 Hz and mark at 1800 Hz for reception on USB. If LSB is used, the frequencies will be inverted. Since the space frequency is fixed for USB reception, the mark filter should be tunable.

table 1. VS-1 frequency relationships.

	frequency (Hz)	
	170 shift	850 shift
audio space	950	950
audio mark	1120	1800
oscillator	9880	9880
space filter	10,830	10,830
mark filter	11,000	11,680

We have defined 11,000 Hz to 11,680 Hz for mark frequencies and 10,830 Hz or 11,850 Hz for space. At this point I chose 10,830 Hz so that (a) 170 Hz would fall at the maximum (counterclockwise) setting of C6, and (b) the dial could be calibrated clockwise in increasing shift frequency. The local oscillator must be tuned to the heterodyned space frequency of 10,830 Hz minus the audio space frequency of 950 Hz for a resultant

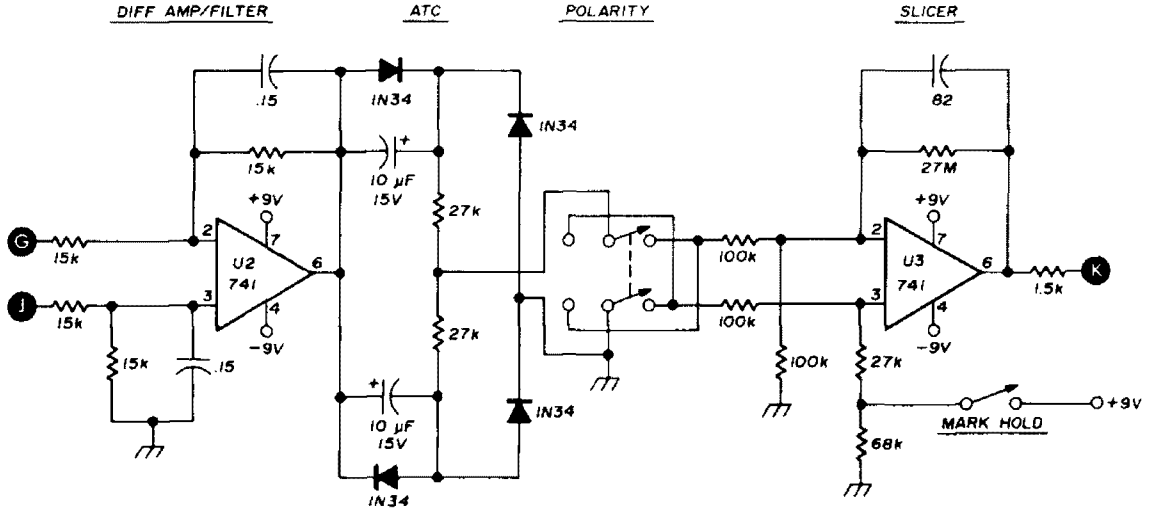


fig. 3. Schematic of the automatic threshold corrector (ATC) and slicer.

of 9880 Hz. Table 1 should make things a little less confusing.

oscillator tuning. The frequency conversion process has been described in some detail, because oscillator tuning must match the receiver bandpass characteristics, and some adjustments may be needed. Also, the oscillator can be moved in frequency to match the more usual tones of 2125 and 2975 Hz for 850-Hz shift if desired. An understanding of the conversion process is necessary to customize this portion of the VS-1. Oscillator tuning can be by means of a front panel control if several different audio frequencies must be used. The filter frequencies may also be increased if a wider range of shifts must be accommodated, but selectivity will suffer.

detectors, ATC, and slicer

Source followers Q5 and Q9 provide stiff drive sources for the detectors while presenting negligible load to the filters **(fig. 3).** Mark and space detectors are identical voltage doublers using germanium diodes for low voltage drop. The outputs are combined in differential amplifier U2, which also provides some filtering. Filter time constants allow transmission rates to 100 wpm. ATC circuitry following the amplifier is from the ST-6, as is the polarity reversing switch. The slicer, U3, is another differential stage, which further squares the keying pulses and adds more filtering. It also has a mark hold input to lock up the keyer for transmitting. In my equipment, this switch is located next to the machine keyboard, and another pole is used to actuate the transmitter. Three positions are used: "Print," "Mark-hold," and "Transmit."

keyer and FSK

Transistor Q10 **(fig. 4)** drives the parallel-connected magnet circuit at the 60-mA level. The 20-ohm resistor in the emitter return provides a signal to the FSK circuit to detect mark current when the keyboard is actuated for transmitting. The mark signal is sufficient to drive U4, the FSK stage, sharply to zero from its "on" biased state. On space signals, it snaps back on to full output.

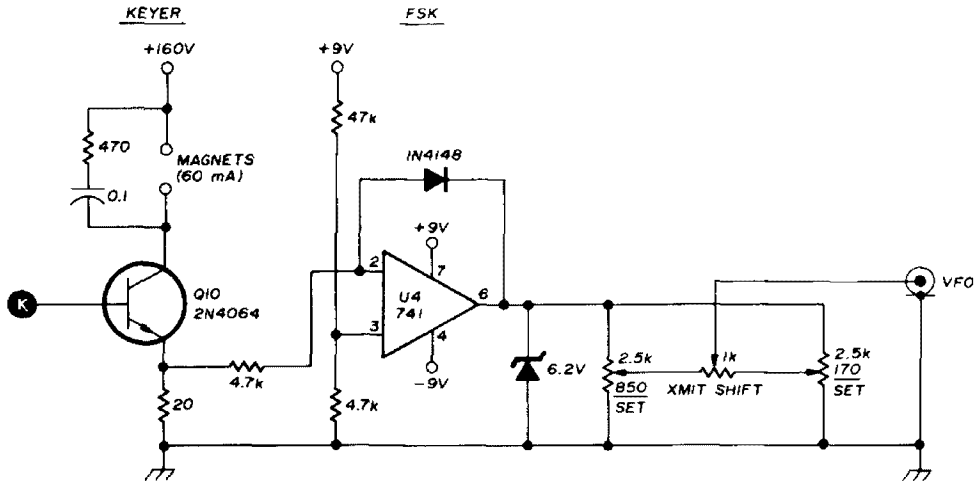


fig. 4. Circuit for the keyer and frequency-shift keyer used in the VS-1 RTTY terminal unit.

FSK control is simple but effective. The positive output voltage of U4 is clipped at +6.2 V nominal by the zener, and the feedback diode limits negative excursions to a negligible value. Thus, we have established two regulated voltages: +6.2 V for space and zero volts for mark. The two divider pots set the end points for the shift pot, so that a continuously variable positive space voltage can be supplied to the transmitter vfo. Note that the shift voltages are not affected by loop current or contact resistance in the keyboard.

Any of the usual transmitter FSK schemes can be used as long as the voltage requirement does not exceed 6.2 V. Mark and space can be inverted by simply inverting the inputs to U4. Output polarity can be reversed by reversing the feedback and zener diodes.

Keying turned out to be simple on my SB-401. The shift pot is connected across the vfo sideband selector diode. The diode is supplied from a high-resistance circuit, and the low output resistance of the shift pot shorts out the normal bias and sets a new level in "transmit." The actual shift depends slightly on vfo tuning, but it is essentially constant over the usual RTTY frequencies, so the shift pot can be calibrated directly in shift frequency from 170 to 850 Hz. The control shown in the photos is a standard pot equipped with a homebrew stop at 180 degrees.

tuning indicator

AC outputs from the filters are isolated by source followers Q3 and Q7, which drive deflection amplifiers Q4 and Q8 (see fig. 5). Separate isolators are used, because the waveforms at the sources of Q5 and Q6 are affected by detector loading. The 2N5550 transistors will deliver about 100 V p-p, which is enough to drive most small CR tubes. Both deflection channels are direct coupled to eliminate trace jumping from noise or overload. The variable bypass circuits in the emitters provide gain control independent of positioning. My VS-1 uses a 1EP1 CRT, but only because I found one at a hamfest for five bucks.

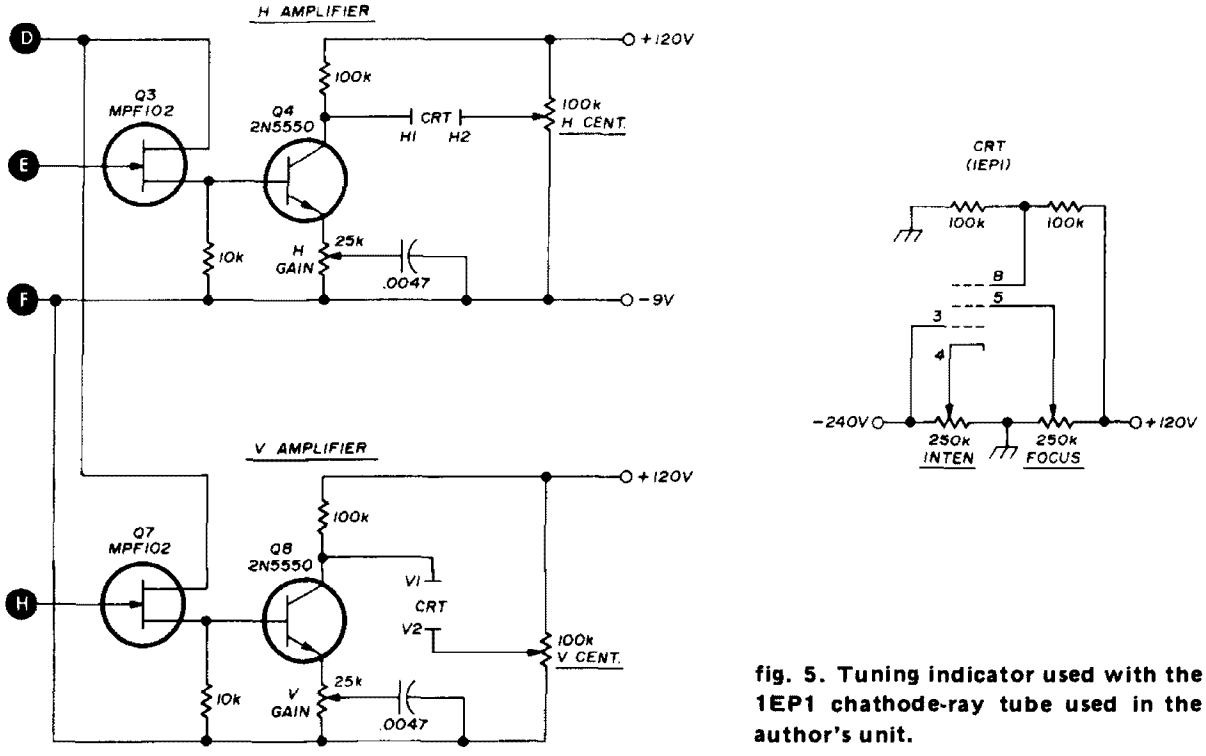


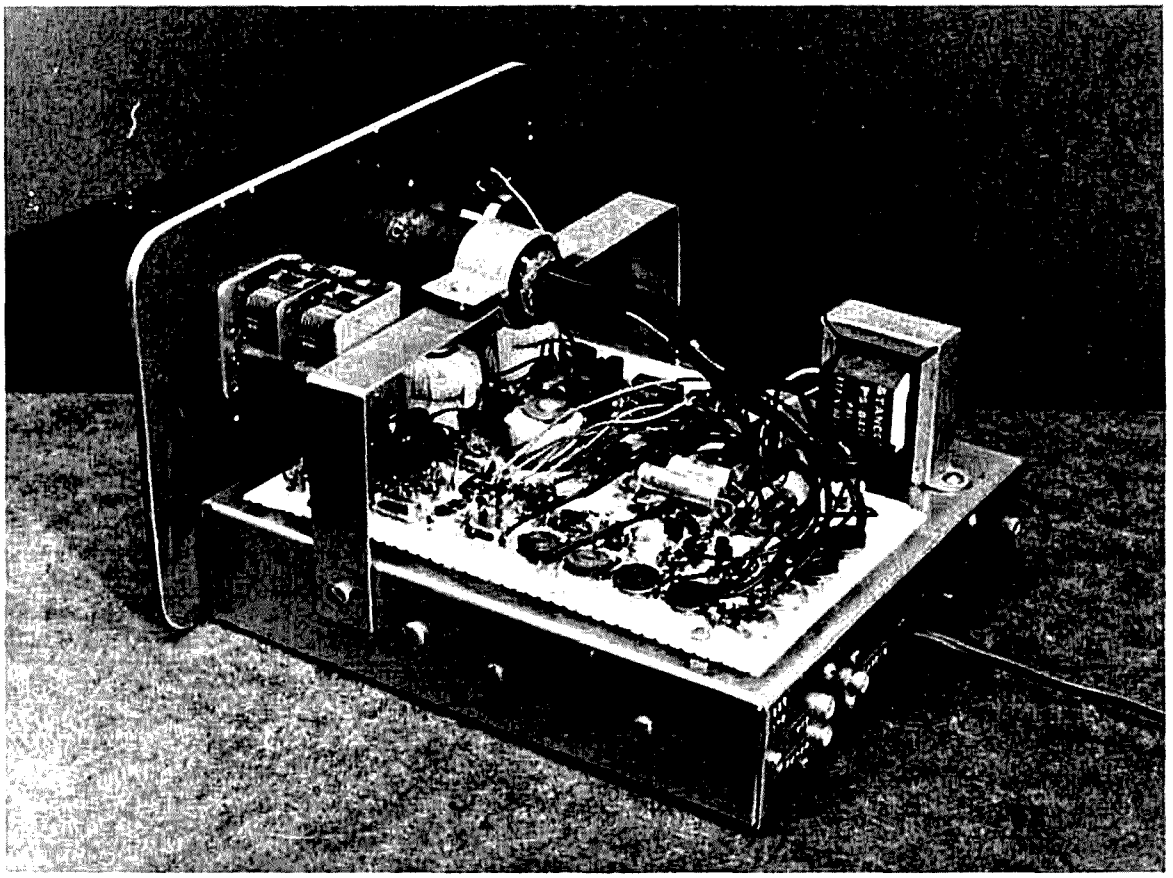
fig. 5. Tuning indicator used with the 1EP1 chathode-ray tube used in the author's unit.

These tubes are too expensive to buy new. Any of the small oscilloscope CR tubes can be used with equal success.

power supplies

Power for the dc loop and low-level stages is derived from a Stancor PS8416 transformer (fig. 6). Although rated for only 25 mA on the dc output, the lack of appreciable loading on the 6.3-V winding keeps transformer temperature reasonable. The 2-μF capacitor provides a higher inrush current on mark signals and allows the average mark hold current to be reduced to 50 mA without degrading print quality.

Low voltages are generated by a pair of voltage doublers and 9.1-V zener diodes. There is nothing mysterious about this arrangement; any low-ripple power supply would work as well. A pair of 6.3-V filament transformers provide an isolated supply for the CRT heater and dc supplies for the intensity, focus, and deflection circuits. Here again, any standard configuration[8] could be used. However, the loop supply should not be used for accelerating or deflection voltage supplies, because the regulation causes trace problems.



Construction of the VS-1 RTTY terminal unit. All circuits except power supply and keyer are mounted on a section of perforated circuit board. Power supply components are mounted under the chassis.

construction and components

Few of the parts in this unit are at all critical, and the mechanical layout is not too important. Common-sense precautions should be taken to isolate filter input and output circuits, but no special bypassing or decoupling is required for stability. Leads to the mark tuning capacitor should be reasonably isolated to minimize coupling around the 18-pF capacitor. If a CRT tuning indicator is used, the tube should be kept away from power transformers.

Except for U1, which handles 10+ kHz signals, all op-amps can be garden-variety 741 types. These ICs are inexpensive and are exceptionally immune to damage. Dual units are available for about $1.00 in TO-type packages from a number of *ham radio* advertisers. Type 709 op-amps could be used, but they are more subject to latchup, require compensation, and will not tolerate output shorts.

Transistors and fets are also garden-variety types except for the high-voltage deflection and keyer units. The fets hang between the power-supply rails, so they should have voltage ratings of at least 20 V. Nearly any npn transistor can be used in place of the 2N3416s. Silicon diodes are low-leakage computer types, and the 1N4148 is suggested for those who must buy them. Power diodes can be any silicon types rated at least 0.5 A and of voltage ratings listed on the schematic.

Germanium diodes can be any rf type but should be checked for high leakage. Capacitor types are unimportant except for C1 through C6, which must be low leakage types such as mylar, polystyrene, or polycarbonate. Trim capacitors can be these types or mica. The total capacitance required at C1 through C5 is 0.002 to 0.003 μF.

Once the filters have been tuned, the oscillator coil should be tuned to roughly 9880 Hz. Next, run a steady signal through the receiver and adjust tuning until two incoming frequencies, 170 Hz apart, are centered in the narrow passband. Trim the local oscillator frequency in the VS-1 to place the lower audio frequency at the peak of the space filter.

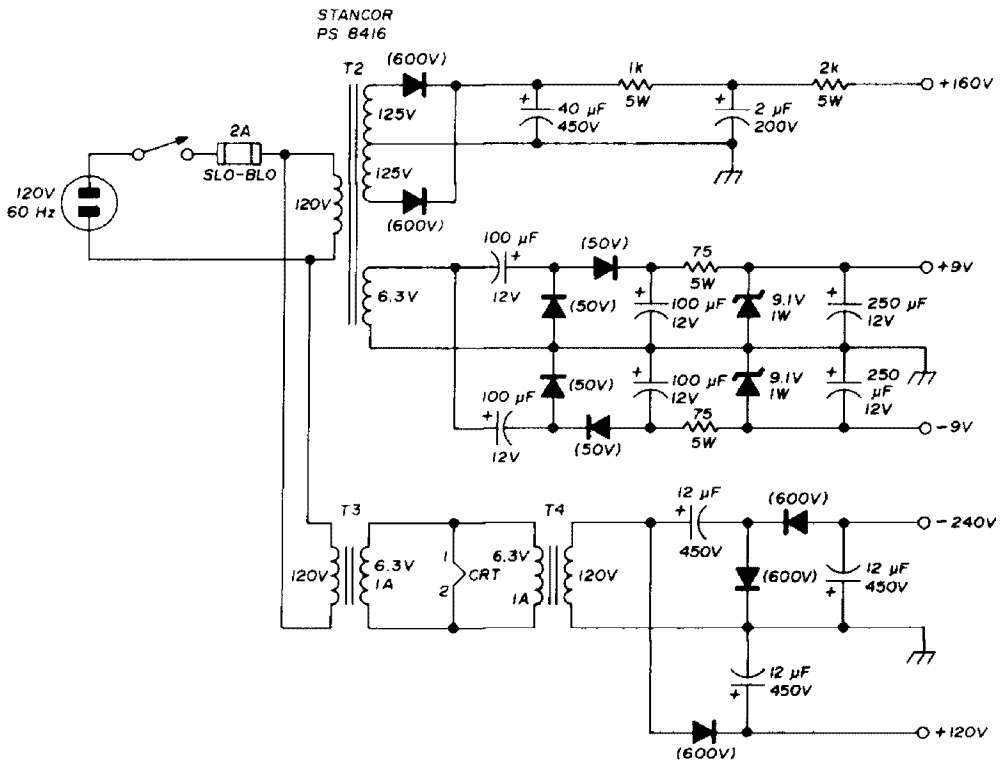


fig. 6. Power supply for the complete VS-1 RTTY terminal unit. Required PIV of diodes is indicated in parenthesis.

adjustments

After construction and wiring are completed, the usual dc voltage checks should be made. Assuming things look normal, the filters may then be tuned. A precise frequency source is not needed, since any errors in alignment will be compensated by the oscillator alignment.

Set the alignment generator to 10,830 Hz, add temporary 82-pF capacitors across L4 and L5, set C6 to maximum, and tune all sections for maximum output from Q5 or Q9. The unit may be fine tuned either by capacitor selection and trimming or by pruning turns. Remove padders, and filter alignment is complete. The variable capacitance of C6 provides the required frequency offsets.

It should now be possible to adjust C6 to place both space and mark signals at the peaks of their respective filters for any shift frequency from 170 Hz to 850 Hz. If a number of different receivers or different audio tones are to be used, a variable tuning capacitor can be used. A 730-pF unit will provide enough range.

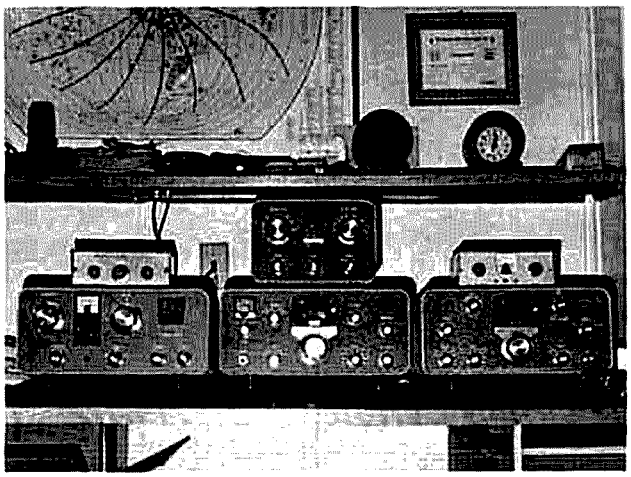
modifications

Motor control and automatic mark hold circuits have been deliberately omitted from this unit, but they undoubtedly will be required by some operators. The tail-end circuitry of the VS-1 is sufficiently similar to the ST-6 so that the same circuit techniques are directly applicable. The power supply may have to

be beefed up to handle the additional load, however.

Some type of CRT display is definitely recommended, since the filters are much too sharp for hit-or-miss tuning. It may be possible to use one of the dual magic eye tubes, but it's doubtful if meter schemes are feasible.

The FSK system shown is by no means the only one that will work. Any of the standard techniques can be used for relay operation directly from the keyboard.



Station at W3VF. The design of the VS-1 is similar to the Heathkit transmitter and receiver used by the author.

For those interested in a hybrid scheme, the 741 at U4 will drive a high-sensitivity reed relay directly with a 12-Vdc coil. This will allow the VS-1 to be used with existing FSK installations. About 10-mA of current is available.

operation

Reception with the VS-1 requires accurate tuning and a reasonably stable receiver. The receiver is tuned to peak the space signal on the appropriate CRT axis, and the shift tuning is then varied to peak the mark signal. Selectivity may be a little scary at first, but only a poor receiver or a drifting transmitter will cause problems. The variations in shift are rather interesting to watch. Advantage should always be taken of the highest degree of receiver selectivity that will pass both space and mark signals.

performance

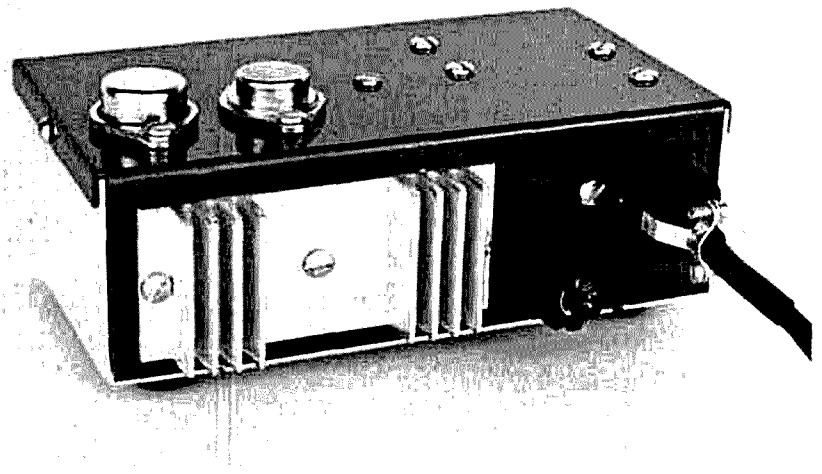
The VS-1 performs well. Filter selectivity is sharp enough to present a clean cross pattern on 170-Hz shift, and the print holds up until signals disappear into the noise. The ATC circuit works well at machine speed but can't cope with keyboard sending under poor conditions with severe signal fading, particularly for loss of the mark signal. A manual offset on the slicer would be required for this condition. Performance in interference is very good, and the improvement in print from using sharp receiver selectivity on 170-Hz shift is striking.

Despite the many words written about limiter/discriminator versus limiterless two-tone reception, the VS-1 shows little difference in print quality as the input gain is varied. Even under poor signal conditions, the limiter can be kept saturated with no increase in print errors. This condition is probably due to the narrow filters, but I'll leave that for others to comment on. The unit is a pleasure to operate, and it now occupies a permanent spot in my station. I hope this article serves to stimulate further work toward improvements.

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ham radio



medium current polarity inverter

Complete construction details for a polarity converter which will provide a negative voltage supply from a positive source

J.R. Laughlin, San Jacinto College, Box 200, Pasadena, Texas 77505

Now you can go mobile with that favorite circuit that has been "grounded" for lack of negative power supply. An increasing number of semiconductor circuits and devices are appearing that require both a positive and a negative power supply for proper operation. For ac-powered installations this is no problem. However, in the automobile or for applications where only storage batteries are available a real problem can develop in supplying the needed negative power.

Often the designer can resort to trick circuitry in order to operate from a positive source. More often, especially in the case of ready made circuits, nothing within reason can be done. The polarity-inverter shown in fig. 1 offers a convenient solution to this problem by supplying a negative potential with power enough to suit most applications. This circuit operates with great efficiency and is virtually noiseless due to its high frequency of operation and lack of the usual power transformer. Filtering is a snap at this high frequency, requiring only nominal values of capacitance to do the job.

The output voltage is well regulated and is adjustable, and the 2-ampere maximum provides up to 22 watts of negative power from your auto battery.

circuit operation

A block diagram of the polarity inverter is shown in fig. 1. The square-wave generator is a stable unijunction transistor oscillator which drives a transistor (Q2) to produce the square wave. Charge time for C2 is controlled by the value of R1 (see fig. 2). Discharge time for this capacitor is controlled by R2.

While C2 is charging, Q2 is saturated; during the discharge time of C2, Q2 is turned off. By proper choice of values for R1 and R2, the output of Q2 will be a symmetrical square wave. Transistor Q3 buffers this output and furnishes driving power for Q4 and Q7. Transistor Q4 inverts the signal from Q3 and supplies base drive to Q5, an emitter follower, which has the capability of driving Q6 into complete saturation.

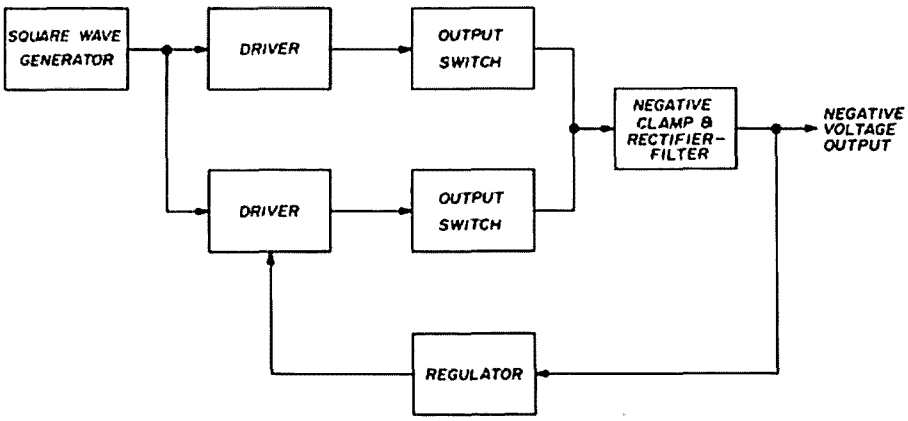


fig. 1. Block diagram of the polarity inverter. With input from an automobile battery, this unit will furnish up to 200 mA at -13 volts, and up to 2 amps at -10 volts.

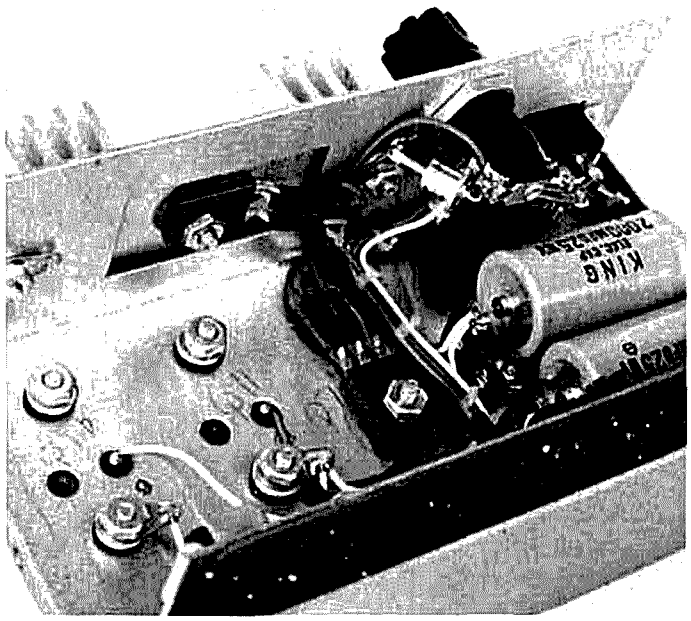
Transistors Q7 and Q6 are connected to form a complementary-symmetry output stage. These output switches are connected in a manner that allows them to be operated in a saturated mode and are driven hard to insure low V_{sat} when they are turned on. This minimizes voltage losses across the transistors and permits the highest possible output voltage.

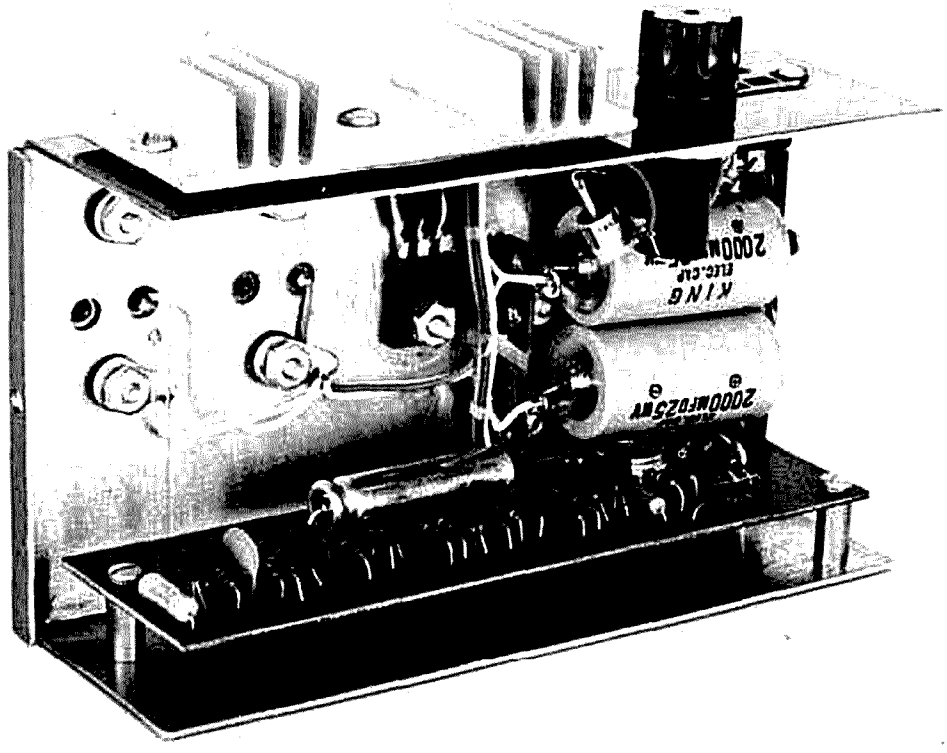
The circuit configuration of transistors Q3, Q4 and Q5 insures that only one of the two output transistors can be turned on at a time. This arrangement is similar to the standard audio amplifier output circuits found in many stereo sets.

As Q7 and Q6 are alternately switched on and off, a square wave that alternates from ground to nearly battery potential is applied to the positive end of C3. Transistors Q10 and Q11 are connected as diode rectifiers. These are germanium transistors and their diode junctions can handle a lot of current with low voltage drop. Their low voltage drop prevents loss of output voltage that would have resulted if silicon diodes had been used.

Transistor Q11 works in conjunction with C3 to clamp the square wave output of Q6 and Q7 negatively. Thus, at the junction of C3 and Q11 and Q10 a square wave is present that varies from approxi-

Power transistor Q6 (2N3055) is mounted on heatsink through an opening in the enclosure. Transistor Q7 (2N2955) uses chassis for heat sinking.





Construction of the solid-state polarity inverter. Majority of components are mounted on printed circuit board (fig. 3) which is installed on half-inch spacers. Heatsink is for power transistor Q6. Transistors Q10 and Q11 are mounted on top of enclosure.

mately ground potential to a negative value almost equal to the positive value of the battery voltage. This negative-going square wave is rectified by Q10 and filtered by C4.

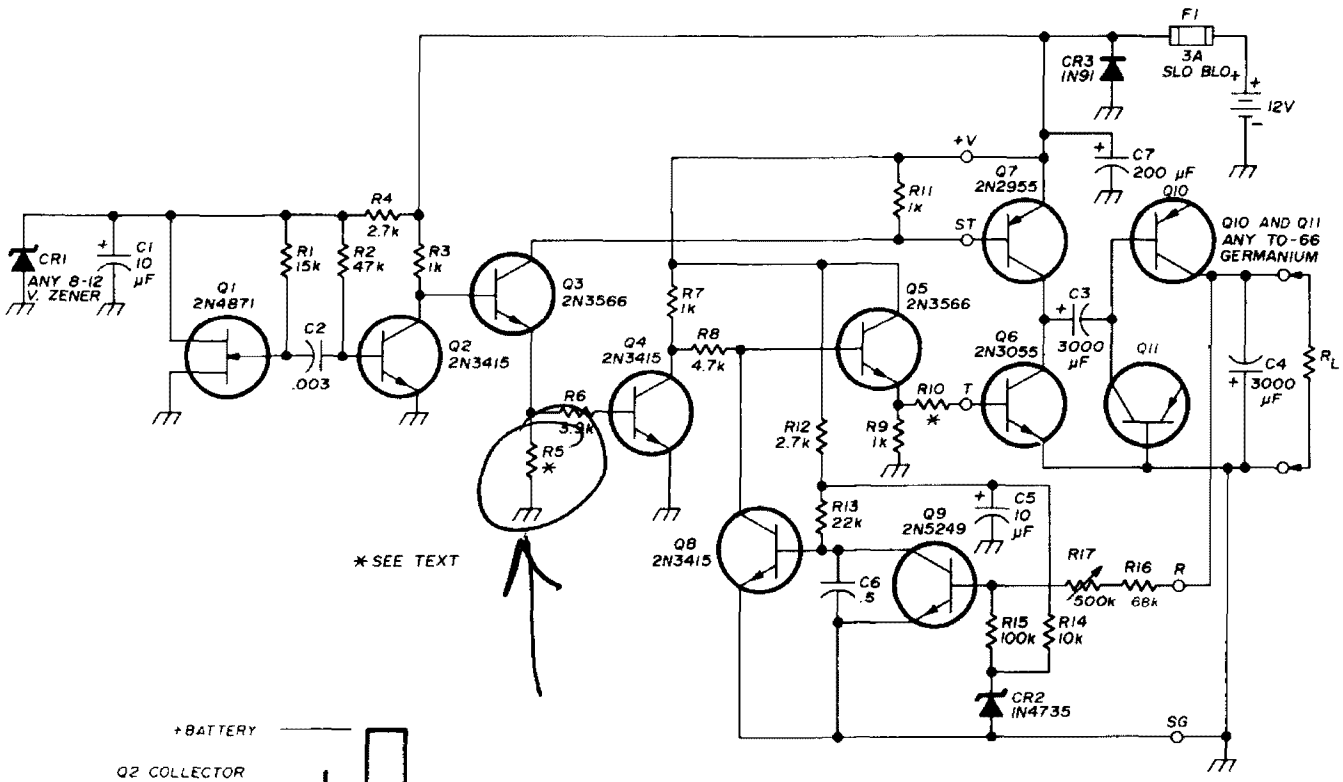
The value of negative voltage across C4 cannot be quite equal to the value of battery voltage because the combined voltage drops across Q6, Q7, and Q10 and Q11 combine to reduce the total output. Naturally, these losses increase with heavier output currents. Table 1 shows available output voltage versus output current.

table 1. Maximum available output voltage vs output current.

output current (amps)	output voltage (-Vdc)
0	14.0
0.2	13.0
0.4	12.7
1.0	11.4
1.5	10.9
2.0	10.0

Regulation of the output voltage is performed by a transistor feedback loop consisting of Q9 and Q8. A zener diode, CR2, provides a stable reference voltage with which to compare the negative output. This comparison is done by Q9. If the negative output voltage attempts to increase, the conduction of Q9 will decrease. The positive-going collector of this transistor will increase the conduction of Q8, thereby preventing the base drive to Q5 from reaching its full positive value.

Reduction of this drive prevents Q6 from being turned on all the way during half of the cycle, resulting in a lower output voltage to the negative clamp circuit and rectifier. Thus, the output voltage is maintained automatically at a level which depends on the relative values of R15 and the series combination of potentiometer R17 and R16. If R16 is



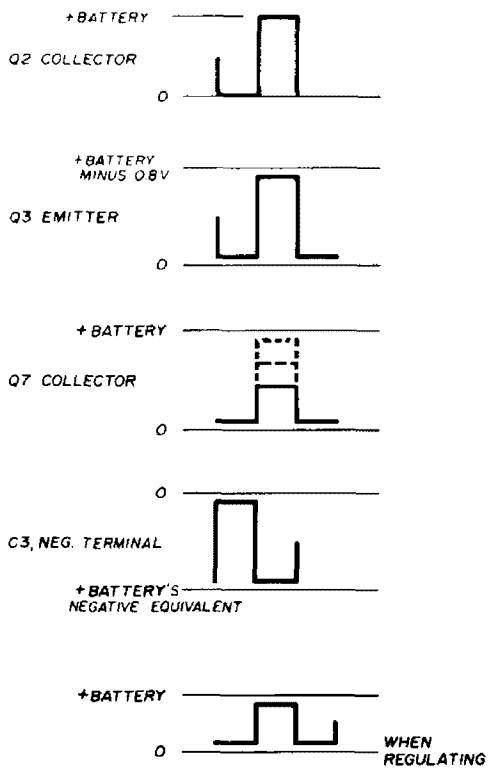


fig. 2. Schematic diagram of the polarity inverter and important waveforms. See text for value of R5.

reduced in value, regulation will occur at a lower output voltage. Table 2 shows approximate output voltage levels for various values of R16 and R17.

Efficiency depends on the difference between the battery supply current and the output current. For example, with a 2-ampere output current the input supply current from the battery will be approximately 2.2 amps. This additional 200 mA of battery current represents a loss of efficiency. Most of this 200 mA is current drive to the bases of Q6 and Q7 through their base-current limiting resistors, R10 and R11. This amount of base drive is not necessary if the inverter is to operate with loads lighter than 2 amps. These resistors may be increased as shown below for the reduced output currents.

maximum output current (amps)	R10 and R11 (ohms)
2	100
1	240
0.5	330
0.25	680

Increasing these resistor values will substantially increase the operating efficiency for lower operating output currents.

construction

It is suggested that you use the circuit board shown in **fig. 3.** This can be

mounted into an aluminum box along with the other components. Heat sinking is not critical for any of the transistors except Q6. This transistor can dissipate quite a lot of power when delivering heavy output current at low output voltage level.

Diode CR3 is not a necessary part of the circuit but was included after the original unit was completed and connected to the battery in reverse! This diode will cause the fuse to open up instead of

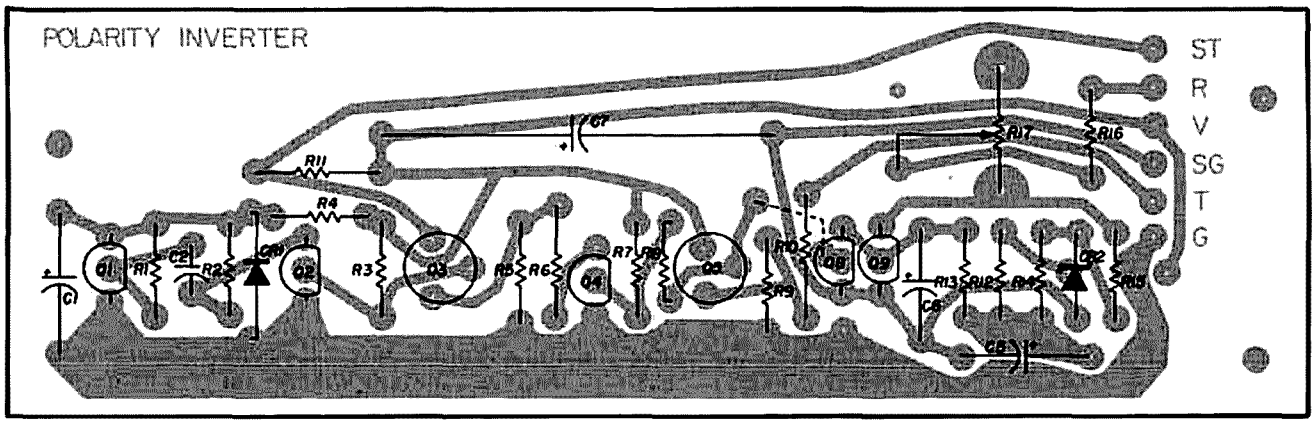


fig. 3. Full-size printed-circuit layout for the polarity inverter.

one or more output transistors if this should happen again. Transistors Q10, Q11 Q6 and Q7 are mounted to the aluminum chassis at widely spaced intervals to facilitate heat dissipation.

checkout

Be certain the fuse is in place before connecting the unit to a battery. Do not bypass the fuse during checkout. It may save a relatively expensive transistor

table 2. Output voltage vs total resistance of R16 and R17 (battery maintained at +14.2 Vdc).

R16 + R17 (ohms)	output voltage (-Vdc)
220k	10
180k	8
160k	7
135k	6
110k	5
96k	4
70k	3

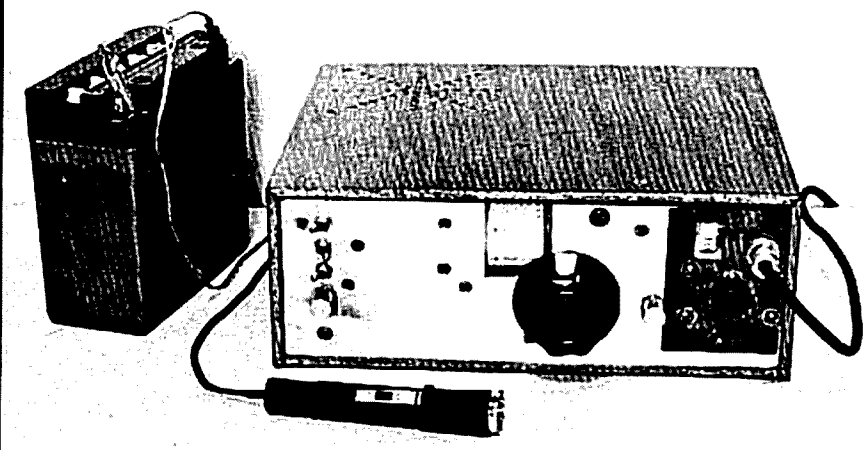
should there be a wiring error. An ammeter in the battery line should indicate approximately 200 mA with the unit turned on and unloaded. This current will be less if higher values of R10 and R11 are used, as discussed earlier.

It is a good proceedure to start at transistor Q1 with an oscilloscope and check each waveform as shown on the schematic. Verify the amplitude of each waveform. If all waveforms are normal, connect a 5-ohm power resistor across the output terminals. With a 5-ohm load, an output voltage of ten volts represents full loading of the inverter. Of course, if you plan to operate at lower output current levels, a higher value resistor may be used for loading the output during the checkout.

Rotation of potentiometer R17 should produce a variation in the output voltage. Alternately connecting and disconnecting the load resistor will verify the regulator action. It should be noted that the maximum regulated output voltage cannot exceed that voltage shown in table 1 for the various values of peak output currents shown.

For example, if your audio power amplifier draws from 0.05 to 1.5 amps on peaks this inverter can supply 10.9 volts of negative regulated dc. If your requirements are from zero to 0.4 amps, the inverter can deliver -12.7 volts dc.

ham radio



single-band ssb transceiver using the LM373 communications IC

How to use the versatile LM373 and several other ICs to build a compact ssb transceiver for 14 MHz

Charles H. Hill, W5BAA, 2510 Beatty, Houston, Texas 77023

About two years ago a new products announcement in *ham radio* described a linear communications IC, the National Semiconductor LM373. Although I have found the LM373 to be the most versatile IC for the communications field on the market, I have not seen any articles in the amateur publications which have given the LM373 the praise which I feel it deserves.

One of the strongest assets concerning the device is the very comprehensive data which the National Semiconductor Corporation supplies on request. With this data you can use the IC in a variety of signal processing roles with a minimum of brain work. In order to sing the praises of the device, I will describe a 20-meter QRP ssb transceiver which I have built around the LM373. Although sufficient information is provided in this article so that you may homebrew your own version, I would strongly suggest that you obtain the data sheets on the IC from National Semiconductor in order that you may fully appreciate the versatility of the device.*

communications IC

Fig. 1 is a functional outline of the LM373. National bills the device as an a-m/fm/ssb i-f strip; however, it is used for a host of other functions including dsb generators and receiver frontends. The package includes an agc-controlled gain stage, the output of which may be used to drive a crystal, mechanical or LC filter; a fixed-gain stage, which may be driven by one of the forementioned filters; a balanced mixer driven by the fixed gain stage and an agc generator which is matched to the agc controlled stage. In addition to the access points for the filter terminations, access is provided for nulling both the signal and local oscillator ports of the balanced mixer.

*National Semiconductor Corporation, 2975 San Ysidro Way, Santa Clara, California 95051.

transceiver

In the transceiver a common i-f strip is built using U1, an LM373 (see fig. 2). By switching the input signal applied to pin 2 from the output of the receiver frontend to the output of the dsb generator, the local oscillator signal at pin 6 from the 9-MHz bfo to the 5-MHz vfo, and the output at pin 7 from the audio amp to the 14-MHz filter, one LM373 IC acts as both a receiver i-f strip with a built-in product detector and agc system, and transmitter filter and hf mixer with a built-in agc controlled speech compression. Of course, a 9-MHz ssb filter, FL1, placed between pins 9 and 4 provides the necessary filtering. All of the switching is handled by diode signal swtiches.

In addition to i-f duties, the LM373 is also used as a dsb generator. In this case the output of a dynamic microphone is fed directly to pin 4 of U2, the fixed gain

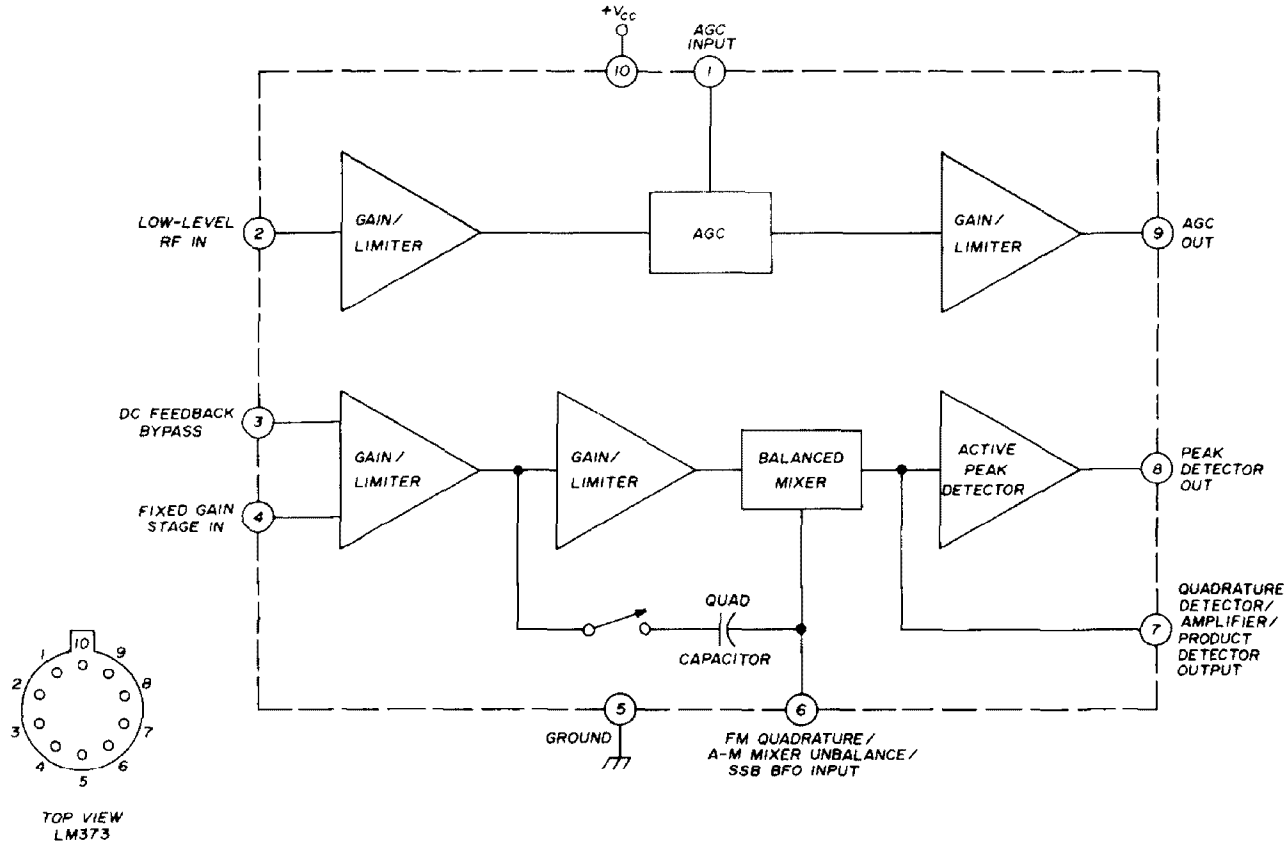
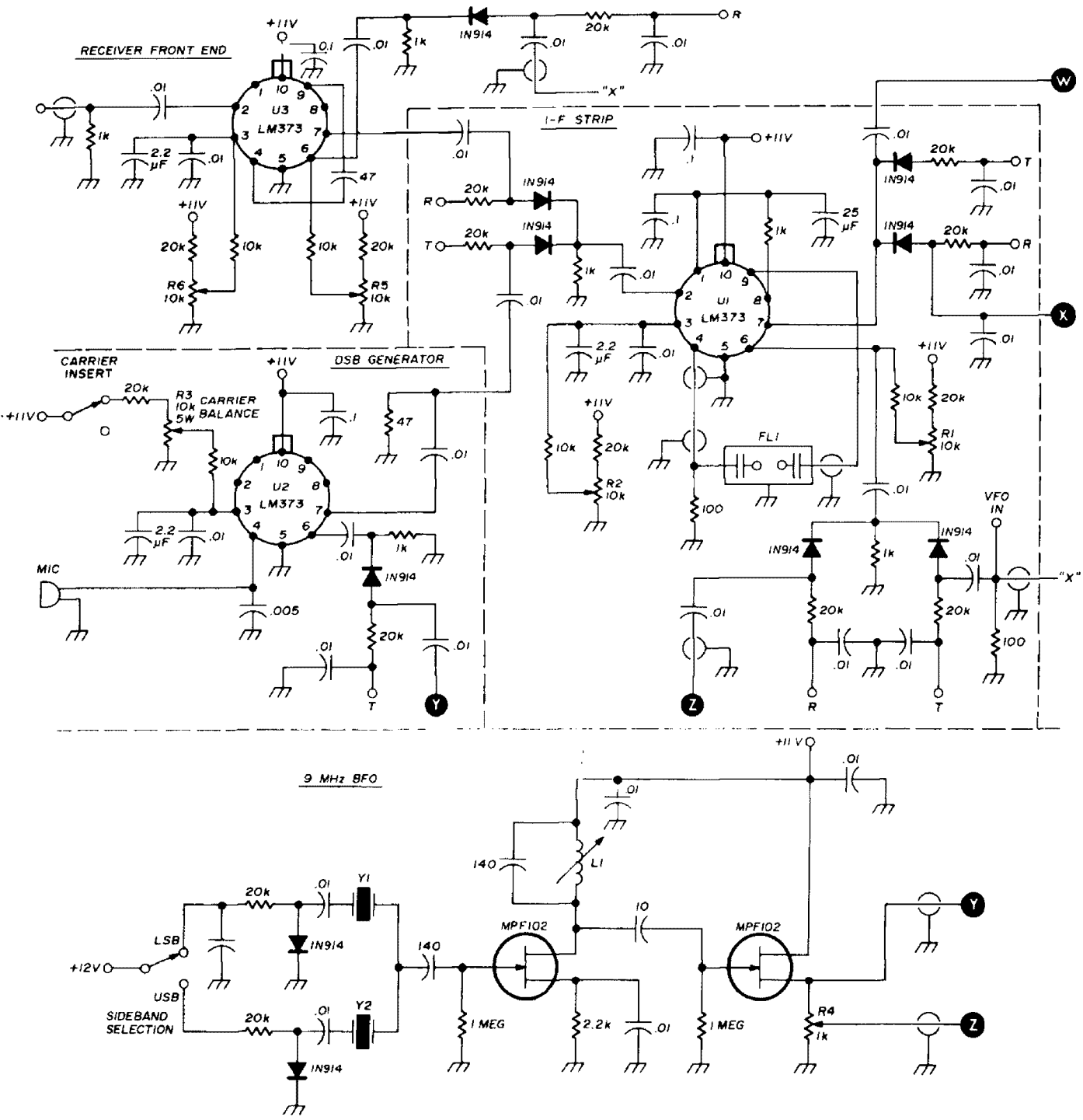


fig. 1. Block diagram of the versatile National Semiconductor LM373 communications IC. This IC can be used in a number of communications circuits, as illustrated in the 14-MHz transceiver circuit shown in fig. 3.

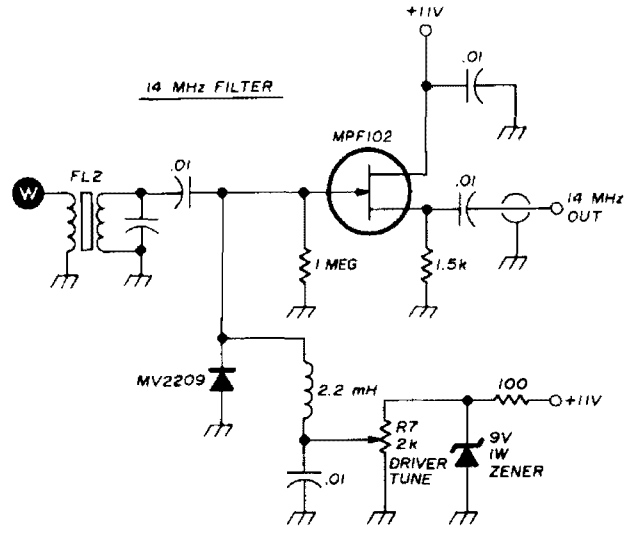


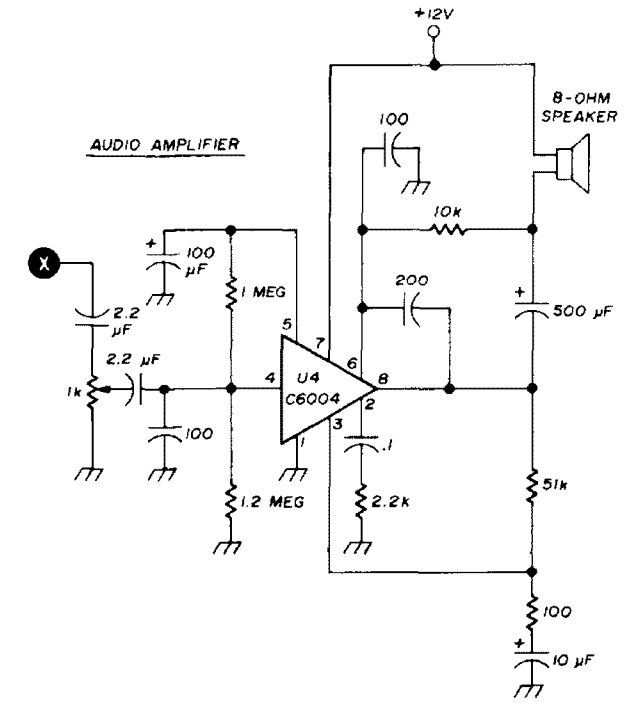
FL1	9-MHz crystal filter (KVG)	L1	24 turns no. 26, spacewound to cover entire length of Miller 4300-2 slug-tuned, ceramic coil form
FL2	Primary, 3 turns no. 26; tuned winding, 26 turns no. 26 on Amidon T-25 SF toroid core	Y1,Y2	Upper and lower sideband crystals for 9-MHz crystal filter (KVG)

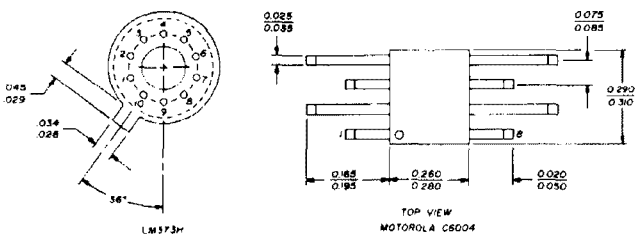
fig. 2. Circuit diagram for the complete 14-MHz CW/ssb transceiver based on the National Semiconductor LM373 communications IC.

stage input. The 9-MHz bfo signal is switched to pin 6 during transmit, and the dsb output is taken at pin 7.

In the transceiver an LM373 is also used as a receiver front end. The input signal is applied to pin 2 of U3, and a 47-pF capacitor couples the output of the agc controlled gain stage (pin 9) to the input of the fixed gain stage (pin 4). The 5-MHz vfo is switched to pin 6 during







receive, and the frontend output is taken from pin 7. All of these LM373 circuits are taken directly from the National Semiconductor application notes.

Audio amplification is provided by the Motorola C6004, U4, which is capable of producing a 1-watt output into an 8-ohm speaker without any transformers. The circuit shown is taken directly from Motorola application notes.

The 14-MHz ssb signal from the i-f strip is separated from the 4-MHz product by the 14-MHz filter, FL2. This resonant circuit is tuned by a Motorola Epicap, MV2209. This allows the driver tune control, a variable resistor, to be remotely mounted on the front panel. An fet drain follower provides a low impedance output from the tuned circuit to the input of the linear amp. The first stage of the linear amplifier is an RCA CA3028A IC in a differential amplifier configuration which drives two 2N2102 emitter followers. These, in turn, drive two 2N2102 transistors in push pull. The final consists of two 2N3553 transistors in class B push pull. The output is transferred by transformer T2 through a coax relay to the 14-MHz input/output filter. This filter serves as both an output filter for the transmitter and an input filter for the receiver.

construction

The transceiver is built on a single piece of copper clad epoxy board with the 5-MHz vfo and linear amplifier mounted separately. The copper is left on to provide the necessary ground plane. Holes are drilled with a no. 60 drill to allow component leads to pass through the board. Ground connections are made directly by soldering to the copper foil. Leads above ground are isolated by reaming away the copper around the holes. The component leads are then hard wired on the non-copper side of the board. All capacitors are rated for 15 volts dc and all resistors are 1/4 watt. All diodes except the Epicap are 1N914 switching diodes. The ssb filter and matching crystals are manufactured by KVG. The only shielding required is around the 9-MHz bfo, around the vfo and between stages of the linear amplifier.

The layout I used is shown in the photograph. An exact layout is not provided here because the actual components I used may be unavailable to you, or may vary somewhat in size. To produce a

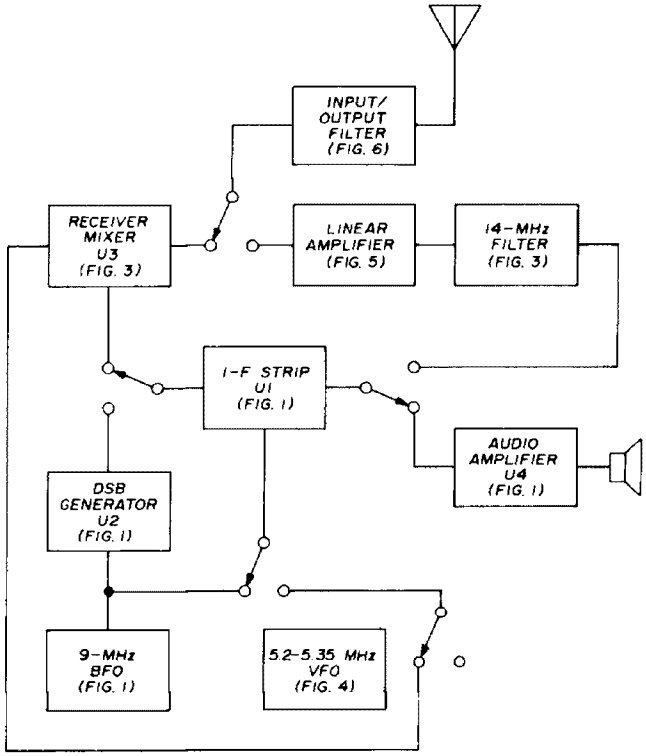


fig. 3. Block diagram of the 14-MHz transceiver in the receive mode. Complete diagram of the transceiver is given in fig. 2.

drilling template, I used ten-squares-per-inch grid paper on which to lay out the components to actual size. It is wise to make all leads as short as possible, and to use shielded lead where indicated in the diagrams. A rigid mechanical frame will insure adequate frequency stability.

tuneup

Since I have a limited amount of test equipment (5-MHz oscilloscope, a Q meter and a crystal calibrator) I had to use a boot-strap method to align the unit. First, check the audio amplifier by touching pin 4 of the C6004 IC with a metal screw driver. The 60-Hz pickup should drive the audio amplifier to good audio output.

Next, the 9-MHz crystal bfo may be brought to life by monitoring its output with an oscilloscope (I am able to see the signal on my 5 MHz scope) or listening to the audio output for a rush of noise while adjusting inductor L1. The i-f strip may be aligned by switching the carrier insert switch to the carrier on position, the unit to transmit, and observing the output of U1 at pin 7 with the scope while adjusting R1 (the signal port null) for a maximum signal output.

Also, it will be necessary to adjust the local-oscillator port null, R2, but first it is necessary to null the carrier from the dsb generator. This is done by switching the carrier insert switch to carrier off and adjusting R3 for a null as observed at pin 7 of U1. Resistor R4 must be adjusted to a threshold point where the null is minimum. Now the i-f local oscillator port may be nulled by adjusting R2 for a null as observed at pin 7 of U1.

Since the signal port adjustment will interact with the local oscillator port null, R1, R2 and R3 must be adjusted in sequence several times to achieve maximum signal output at pin 7 with the carrier on, and for minimum signal with the carrier off.

At this point the i-f strip and the dsb generator are aligned. The receiver front-end may be aligned with or without a signal generator. To align the circuit with

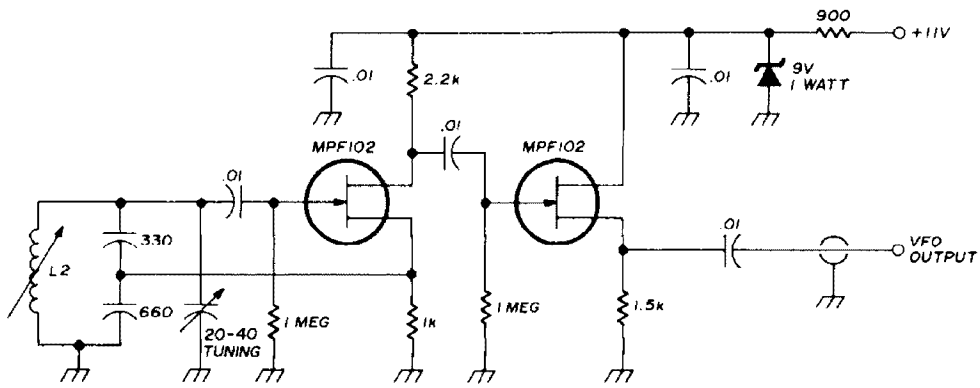
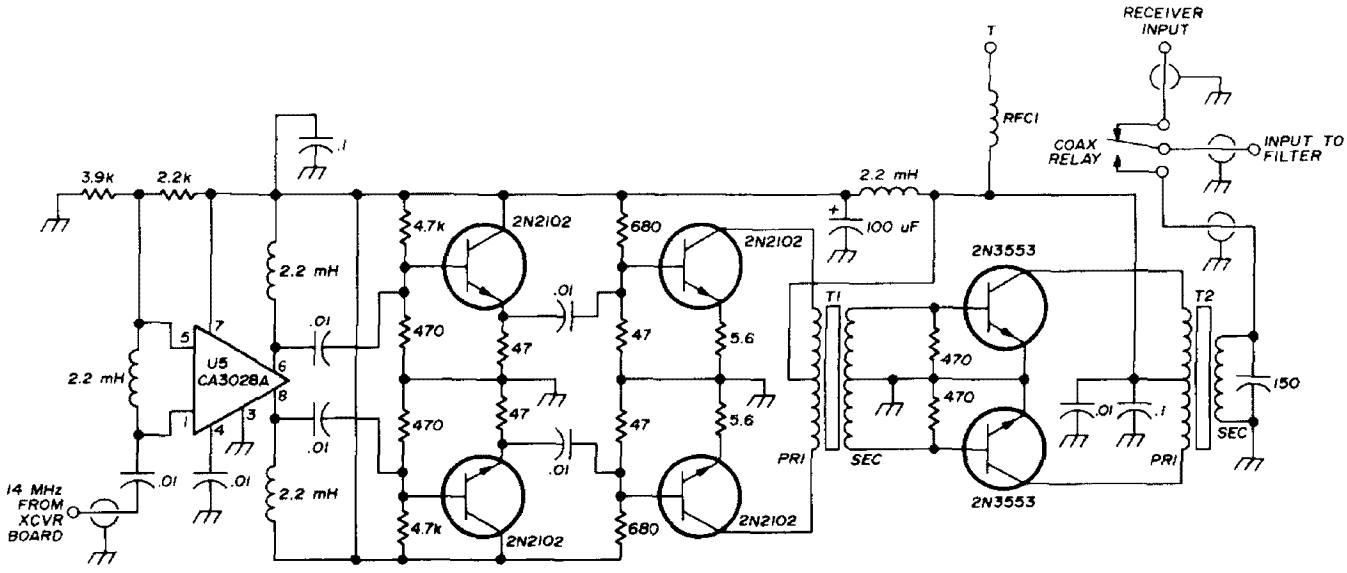


fig. 4. Vfo for the 14-MHz transceiver shown in fig. 3 covers the range from 5.2 to 5.35 MHz. All fixed capacitors are silver-mica units. Inductor L2 is 35 turns no. 26, scramble wound over one-half the length of a Miller 4300-2 ceramic coil form.



RFC1 25 turns no. 16 on Amidon T-50 SF toroid core

T1 Primary, 20 bifilar turns no. 26; secondary, 6 bifilar turns no. 26 on Amidon T-37 SF toroid core

T2 Primary, 18 bifilar turns no. 26; secondary, 26 turns on Amidon T-68 SF toroid core

fig. 5. Solid-state linear amplifier provides up to 5 watts output at 14 MHz.

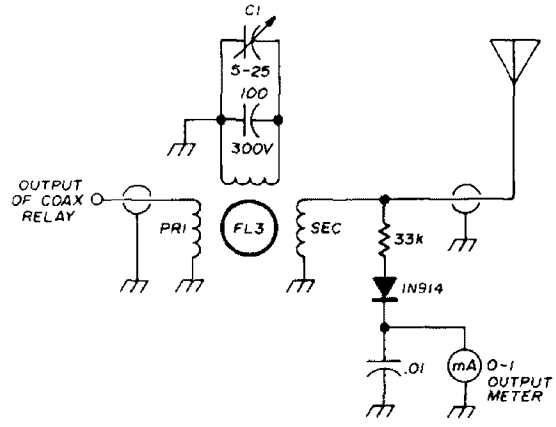
a signal generator, feed the 14-MHz generator output into the antenna terminal and tune L2 until a signal is heard. Next, adjust the signal port null control, R5, for a maximum signal as observed at pin 7 of U1. Also, check for peak tuning of the input/output filter by adjusting C1 for maximum signal at pin 7. If a signal generator is not available, attach an antenna during daylight hours and tune L2 until 20-meter signals can be heard, then adjust R5 and C1 as described above. Now the local oscillator port may be nulled by removing the antenna or signal generator and adjusting R6 for minimum signal at pin 9 of U1. This concludes the receiver frontend alignment.

The linear is a wideband amplifier and requires no alignment. However, if trouble is experienced, bypass the 14-MHz filter and feed the signal from pin 7 of U1 directly into the linear amplifier. This signal will contain a 4-MHz component which can be detected by the scope, thus facilitating normal amplifier troubleshooting procedures. To peak the signals into and out of the linear amplifier, adjust R7, the 14-MHz filter tuning, and C1, the input/output filter tuning, for a maximum signal as indicated on the output meter. This should be done with a 50-ohm antenna or a 50-ohm dummy load attached to the antenna terminal, and the carrier insert switch in the carrier *on* position.

When you have reached this point you can calibrate the transceiver by whatever means available and try it out on the air.

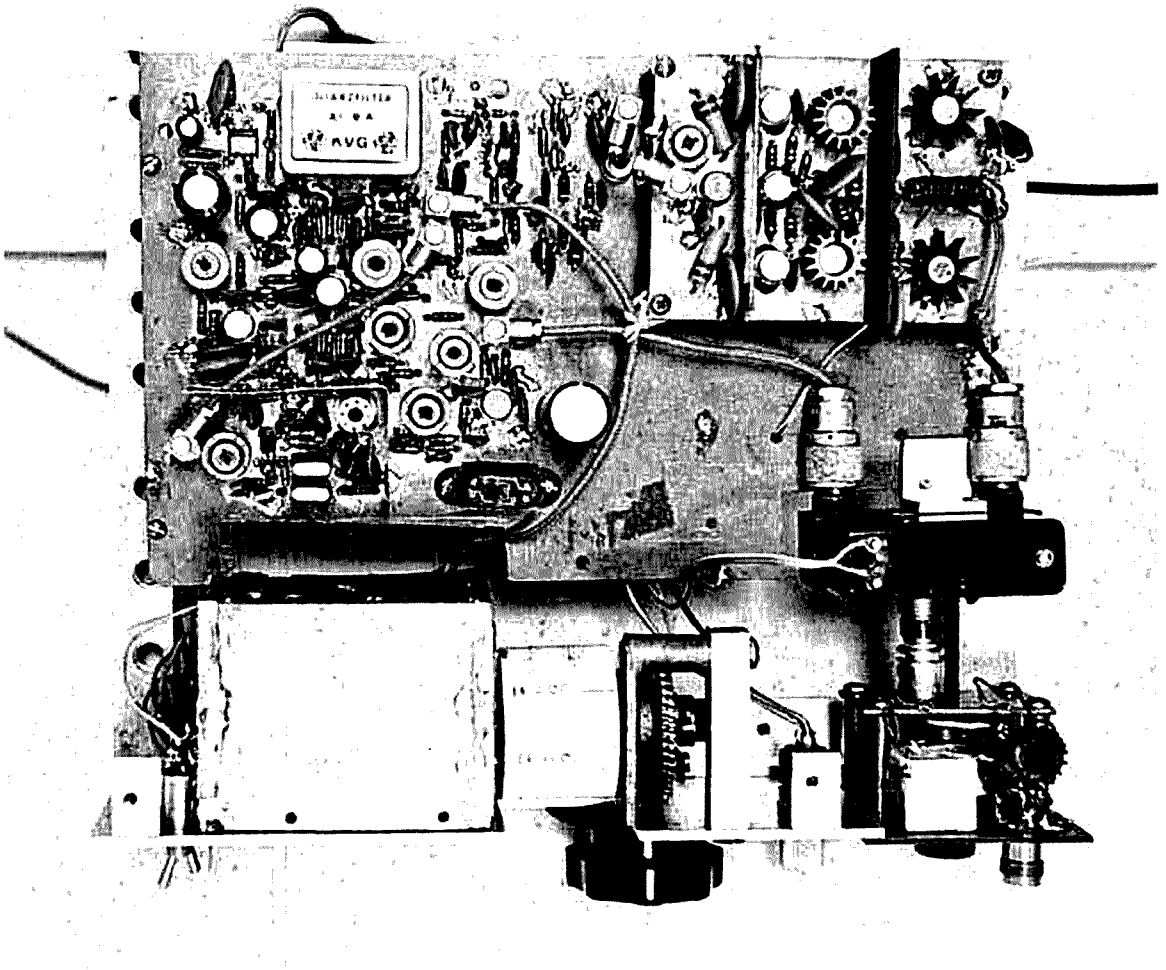
performance

With only 5 watts PEP output a little operator skill is required to communicate.



FL3 Primary, 3 turns no. 16; secondary, 2 turns no. 16; tuned winding, 16 turns no. 16, all on Amidon T-50 SF toroid core

fig. 6. Input/output filter for the 14-MHz CW/ssb transceiver.



Layout of the 14-MHz transceiver, showing the location of the major components.

However, I never cease to be thrilled to announce that I am using QRP after receiving a good signal report. Always get a signal report before telling the other fellow that you are QRP — by some strange phenomenon my signal always goes down after announcing my power level. I must admit that several unsolicited compliments of the audio quality have been made which I attribute to the agc action of U1 during transmit.

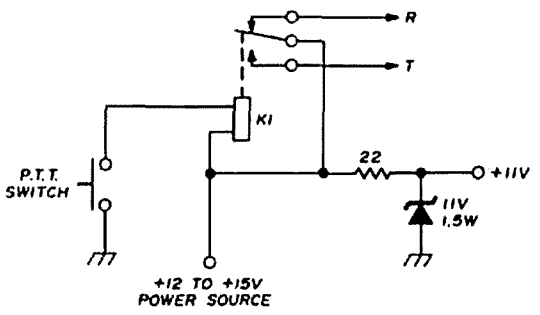


fig. 7. Dc power switching for receive/transmit control on the 14-MHz transceiver.

refinements

If you want to add 75-meter operation, a 4-MHz filter could be switched into the signal path between U1 and the linear amplifier, and a 4-MHz input/output filter switched in between the coax relay and the antenna. Also, a trimmer capacitor would have to be switched into the vfo tank circuit to obtain proper frequency coverage.

The fet oscillator circuits could be replaced by the National LM375 IC, which is a linear IC designed for oscillator/buffer duties. If more power output is desired, a linear amplifier built around the Motorola 2N6367 or 2N6368 might be considered. These transistors are silicon npn devices designed especially for hf ssb service. Motorola provides very good information on the use of these devices from 4 to 30 MHz.

ham radio

single-frequency repeaters for vhf fm

A discussion of single-frequency fm repeaters and how they might be put to use on the vhf amateur bands

George Allen, W2FPP, 4059 Bay Park Drive, Liverpool, New York 13088

A repeater, by definition, is a device used to automatically relay radio, television, telephone or telegraph signals and is used to extend communications over a range not possible with direct communications. The type of repeaters which most people are familiar with are radio or television repeaters which receive on one frequency and transmit on another (fig. 1A). This type of repeater is very common and is the type of repeater which most of the electronics industry has been devoting its time to. However, there is another type of repeater about which very little has been written. This is the single-frequency repeater — a repeater which transmits and receives on the same frequency. This article presents a discussion of the single-frequency repeater.

single-frequency repeater

The single-frequency repeater is broken down into two classes — the delayed type and the simultaneous type. The *delayed* repeater (fig. 2) can operate either on a single channel or use separate frequencies for input and output. This type of repeater records communications on the input channel and re-transmits the recorded communications on the output frequency at a later time. The received signal is not transmitted simultaneously with reception but is *delayed* for later transmission. This type of repeater is useful for data or teletype transmission but is not suitable for voice operation where the receiving station is expected to make an immediate reply.

The *simultaneous* single-frequency repeater uses the same frequency for input and output and appears to transmit the received signal during the period of reception. Comparatively little work has been done on this type of repeater in recent years, and a review of the *Applied Science & Technology Index* back through 1968 shows no references to papers or articles devoted to this type of repeater. Although I have heard rumors that both General Electric and Motorola have built and operated single-frequency repeaters, I haven't been able to find any details.

The simultaneous single-frequency repeater can be further broken down into two sub classes — continuous and time division. The continuous single-frequency repeater (**fig. 3**) transmits and receives continuously on the same frequency. To make this type of repeater work, the input and output must have separate antennas which are completely isolated. Total isolation between input and output must be achieved so that any signal from the transmitter must be below the receiver noise level. From a practical standpoint in the vhf region this is next to impossible to achieve using a single repeater site. Even if this degree of isolation could be achieved, other problems would make this type of repeater impractical. For one thing, transmitted signals would be reflected back from the geographical terrain and any objects in the vicinity of the repeater, so it would be nearly impossible to keep the transmitted signal from getting back into the receiver.

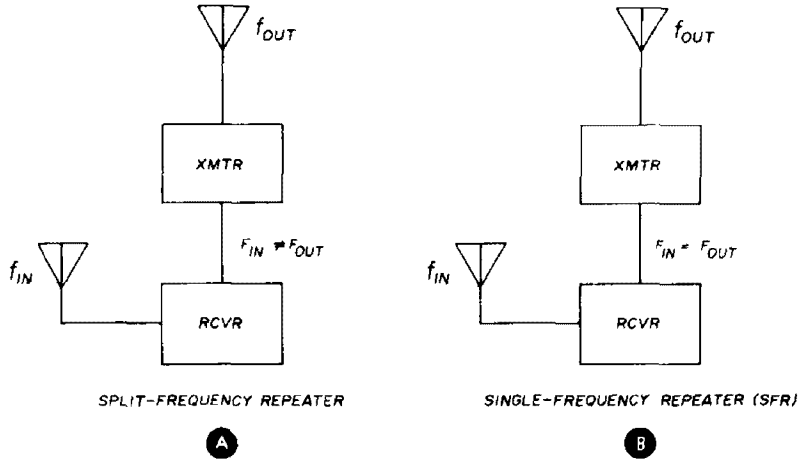


fig. 1. Two basic types of repeaters. The conventional split-frequency repeater shown in (A) transmits on one frequency and receives on another. The single-frequency repeater in (B) transmits and receives on the same frequency.

A more practical approach would be to use a time-division single-frequency repeater as shown in **fig. 4.** This type of repeater is similar to the delayed type in that the input is delayed slightly before transmission. The transmitter and receiver operate alternately, but at such a fast rate that they *appear* to be operating continuously and simultaneously.

Fig. 5A shows a simple plot of the operating period for the time-division SFR. Note that in a given time interval a portion is used for receiving and a portion is used for transmitting. During the receive interval the received information is saved, either by using a delay line or some type of recording device, such as a tape

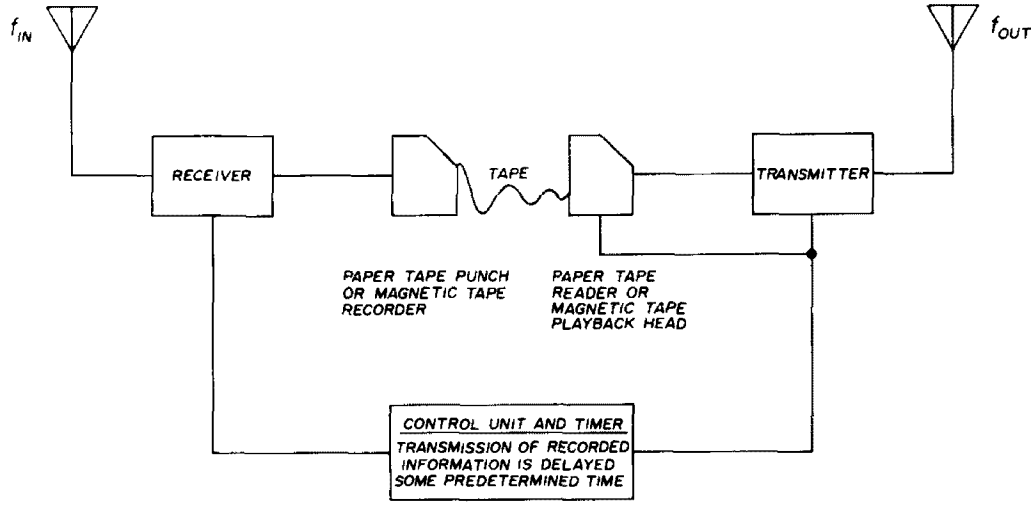


fig. 2. Basic layout of the delayed single-frequency repeater. Input and output frequencies are the same.

recorder. During the transmit interval, the recorded or delayed information is retransmitted.

At this point several problems are introduced into the time-division SFR. First, receivers and transmitters do not turn on and off simultaneously. There is always turn-on and turn-off lag. Thus, after the receiving interval is finished it is necessary to wait for some time period t_{RS} to give the receiver a chance to turn off. This time interval will depend on the type of the receiver and the muting and blanking circuits involved. In regard to the transmitter, it is necessary to wait some time interval t_{TS} before turning on the receiver to prevent the transmitter output from feeding back into the receiver. This is shown in **fig. 5B.**

Another problem arises due to the fact that the receiver is not receiving continuously. There is a significant time period for which information is not received, but lost. This is a sampling mode of operation so the signal to be repeated is not repeated in its entirety, but only samples of the communication are repeated. This problem of lost information can

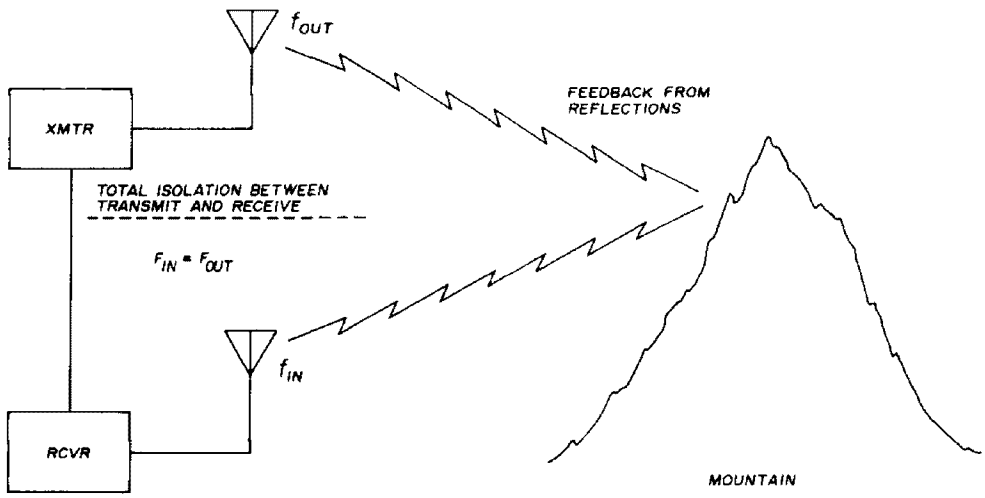


fig. 3. Continuous single-frequency repeater is not practical because any reflections from the transmitter are fed back into the receiver.

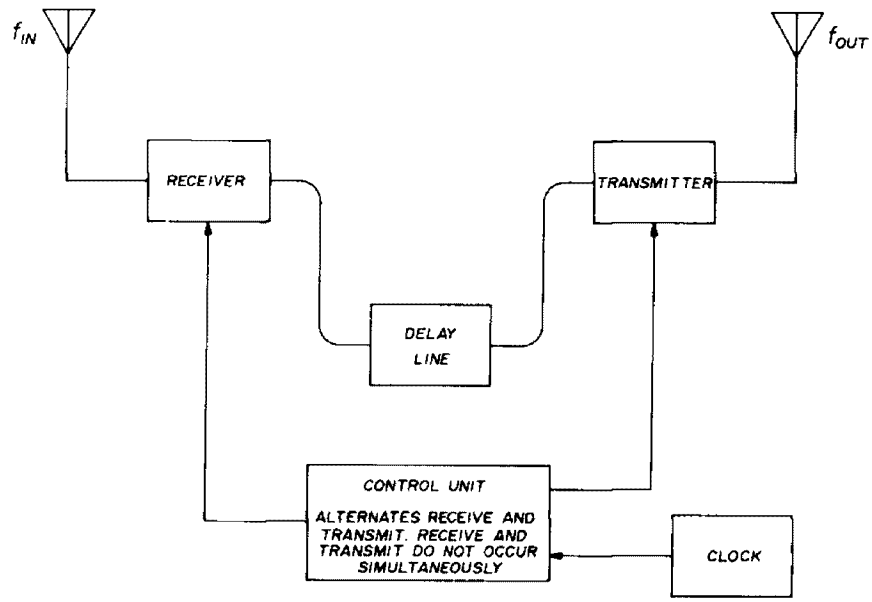


fig. 4. Time-division single-frequency repeater alternates transmission and reception. The repeater appears to be on continuously, but is not.

be solved by choosing the proper sampling interval.

The telephone companies have been using similar sampling techniques for overseas cables, so it is possible to select a sampling interval which will provide minimum loss of intelligibility. This interval, which is rapid enough to retain intelligibility, must be slow enough so that the receive-transmit times are greater than the settling times. If the settling time becomes too large as compared to receive-transmit time, then too much information will be lost and intelligibility will be impared.

Another factor in the choice of time interval is the availability of delay lines. The delay line or recording media must delay the input for a period equal to the receive time plus the receive settling time.

mobile considerations

When attempting to receive a transmission from a time-division SFR a couple of things will be evident. For one, as shown in fig. 6, a noticeable buzz or noise could be apparent at the mobile receiver due to the fact that the transmitter is not transmitting continuously but is being on-off modulated in a pulsed mode. This pulsed modulation will appear as a buzz at the mobile receiver and may possibly override the information being repeated. Some means must be used to eliminate this undesired noise. One possible solution would be to use standard noise blanking techniques. Perhaps a more practical approach would be to use a type of synchronized, muted detector. This type of detector would be similar to the detector used to detect synchronous data transmissions. It would have a time interval sensing circuit that would use the first few time intervals to determine the length of interval that the repeater is using. After this interval has been determined, the detector would anticipate future time intervals and in turn, mute the received audio during non-transmit periods. This would eliminate the received pulse noise.

In regard to the mobile transmitter, it appears that no changes would have to be made. In cases where the received mobile signal is stronger from the other mobile than from the repeater, the synchronized, muting detector would capture the mobile signal and operation would be as if there were no repeater. It might be

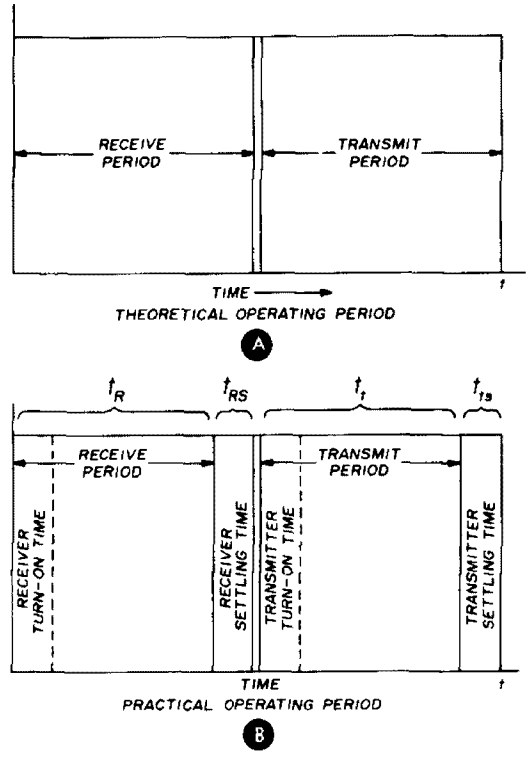


fig. 5. Operating period of the time-division single-frequency repeater.

mentioned that with this type of system, de-sensing would not occur when mobiles that were using the repeater were close to each other.

While all of the above considerations may appear to present a horrendous picture of impracticality for this type of repeater, you must recognize that each of the concepts are practical from a state-of-the-art standpoint. The only concept not proven is the sum total of all the techniques working together.

experimental single-frequency repeaters

A single-frequency repeater can be

built for experimental purposes provided that some assumptions are made. These preclude any exotic circuitry and emphasize simple compromise approaches. The first assumption is that a short timing cycle would present difficulty due to the complexity of high-speed switching circuitry. Thus, a long timing cycle must be used. Second, delay lines for storage of information from receive to transmit may not be available, so simple means such as tape loop delay lines must be used. Third, commercially available receivers and transmitters must be used to simplify repeater construction and permit use of existing transmit/receive circuitry. Finally, the information to be repeated must be slow in nature so that a signal loss of seconds would not result in a loss of information. It is also assumed that due to the long operating period, noise in the mobile receiver would not be a factor. Thus, this experimental repeater would be used basically to repeat frequency modulated telemetry or on-off tone signals of long duration.

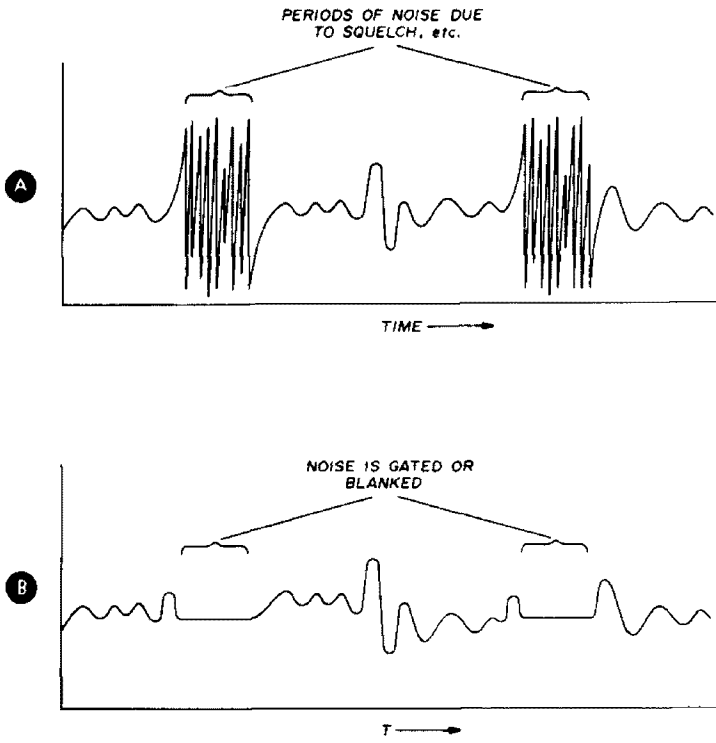


fig. 6. With a time-division single-frequency repeater the operator of a base or mobile station would hear a buzzing sound due to the periods between transmission as shown in (A). This could be solved with noise blanking or gating techniques as discussed in the text (B).

practical single-frequency repeaters

The single-frequency repeater shown in fig. 8 is designed to repeat tone-frequency modulated telemetry. The basic unit consists of a transceiver such as a Standard 826MA, a clock which generates 0.1-second pulses, a slow speed tape recorder with separate playback head, and a simple control unit. The control unit is set up to count pulses and control the repeater in a manner such that the receiver receives for 9 pulses, waits for a 1 pulse settling period, the transmitter transmits for 9 pulses and waits for a 1 pulse settling period. The record head is connected to the receiver and the playback head is connected to the transmitter. The recorder is operated at the slowest possible speed. Note that it may be necessary to move the playback head or provide another playback head at a distance far enough from the record head to provide an approximate one-half second delay.

The single-frequency repeater shown

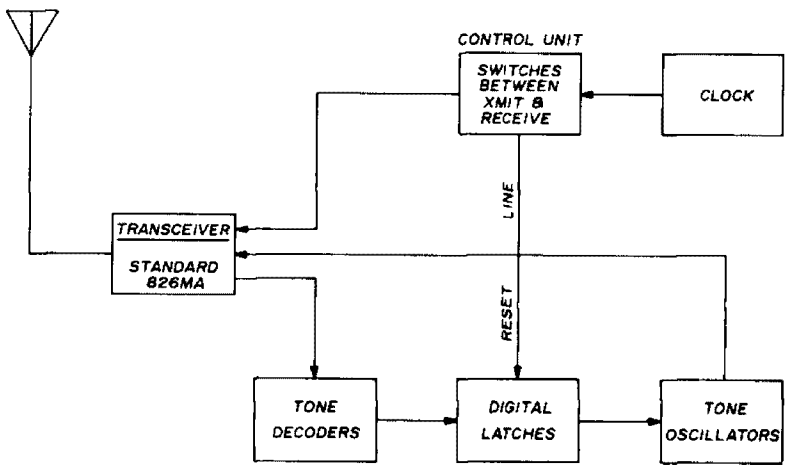


fig. 7. This practical single-frequency repeater is similar to the system shown in fig. 8 except that the tape recorder is replaced by a series of tone filters, tone oscillators and digital latches (see text).

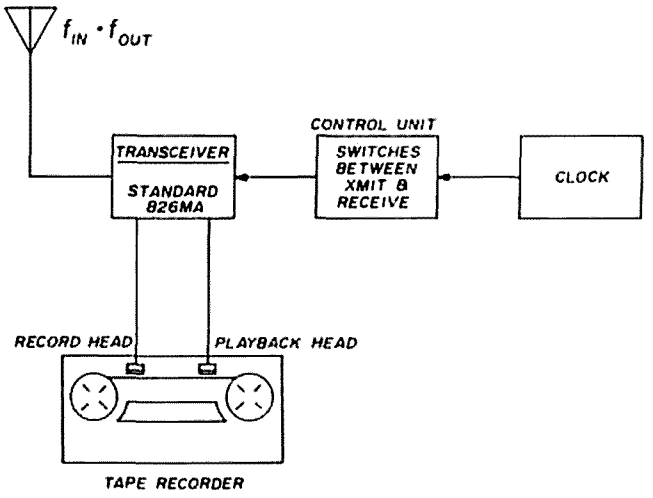


fig. 8. Practical single-frequency repeater. The time delay between record and playback is equal to one-half the operating period.

in fig. 7 is the same as SFR shown in fig. 8 except that it is designed to repeat on-off signaling. The tape recorder delay line is replaced by a series of tone filters, tone oscillators and digital latches. The tone filters determine which tones have been received and set the appropriate digital latches. When transmit time comes, the digital latches cause the proper tones to be re-transmitted. At the end of the transmit period, all latches are reset and the system is ready to receive.

Note that any transceiver may be used in either of the single-frequency repeaters. However, the Standard 826MA was chosen because of its commercial quality and reliability.

summary

The purpose of this article is to present a simple discussion of the single-frequency repeater and to stimulate discussion and experimentation on the subject. It is hoped that enough ideas have been presented to kindle an interest in other experimenters.

In regard to future work, I am presently working on the two practical SFRs as discussed in this article. I have filed a license application with the FCC for this type of repeater and will start on-the-air tests as soon as the license arrives. I expect to report the results of these tests within the next year.

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open-wire impedance-matching baluns

How to design and use transmission line sections for impedance-matching baluns

R. W. Johnson, W6MUR, Post Box 803, Ben Lomond, California 95005

Greater emphasis on the low-frequency bands as the sunspot cycle decreases makes it desirable to more carefully consider the use of wire antennas. In two previous articles on this general subject I have described a two-band matching system using open-wire lines[1] and a means of using lumped capacitances instead of open stubs in a double-stub matching system.[2] In this article I will discuss two convenient wire-line baluns that will also serve as impedance transformers.

Baluns made of coaxial cable, or toroidal transformers, are limited in their impedance transforming capability because you cannot conveniently change the Z_o of the coax or the turns on the transformer. The balun must be used in addition to the matching system. It is relatively easy, however, to make a balun using open-wire lines that will give a much wider choice of matching ratio, and thus combine the matching system and the balun into one unit to do both jobs at once.

In working with transmission lines it is helpful to use the pi equivalent circuit which is given in **fig. 1.** While this may not be physically realizable using lumped constants, except at one frequency, it is nevertheless a valid mathematical model which helps us to more easily visualize what is going on with transmission lines than if we try to work exclusively with the transmission line equations or graphical devices such as the Smith Chart. **Fig. 1B** shows the true balanced configuration; if you are unconcerned with balance to ground, simply make the top series reactance $jZ_o \sin \theta$ and eliminate the bottom one, as in **fig. 1C.**

Looking at **fig. 1** you can see that a transmission line without any connected reactances is always symmetrical; that is, the shunt arms are equal to each other. There is only one symmetrical pi network that will exhibit transformer action between two pure resistances, and that is the one in which each reactance is equal to $\sqrt{R_{in} R}$, and where the series and shunt reactances are of opposite sign. Thus, in **fig. 1** you must have $\sin \theta = 1/(\tan \theta/2)$. With the aid of trigonometric identities it is easy to show that this requires that θ

be some odd multiple of 90°; e.g., a quarterwave line. The transformation is simply $R_{in} = Z_o^2/R$ for this special case.

In other cases, transformer action between two pure resistances demands that there be some added reactance to one or the other of the shunt arms, so that the network becomes asymmetrical. This is why a stub or lumped reactance of some sort must be added to the general line to achieve a match, although there have been arrangements using an off-resonant antenna as the added reactance on the load side.

to ground. Thus, a coaxial cable can be attached to terminals 1 and 2 and a balanced antenna to terminals 3 and 4; conversely, a grounded antenna may be connected to terminals 3 and 4 and an open-wire line connected to terminals 1 and 2. Note also that with the load disconnected, the input resistance is zero since the bridge is essentially two series-resonant circuits in parallel. This facilitates tuning the network in cases where the reactances are made adjustable.

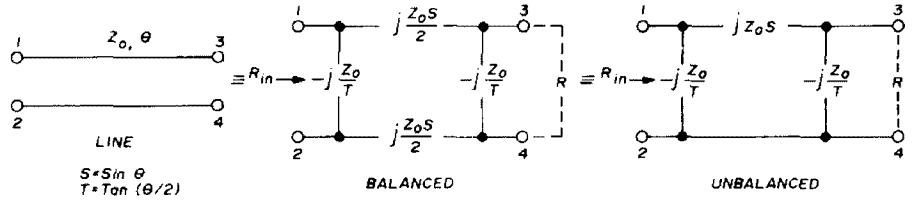


fig. 1. Pi equivalents of the transmission line.

lattice networks

It is also desirable to consider the *lattice* network as an impedance transformer. The lattice is shown in fig. 2 in both conventional and bridge forms. Again, for the symmetrical network there is only one configuration where transformer action takes place between two pure resistances, that being where the shunt and series arms of the lattice network are pure reactances, equal in magnitude and opposite in phase, $X_1 = -X_2 = X$ in fig. 2. The transformation again is simply $R_{in} = X^2/R$ where X is the reactance of each of the lattice arms.

The important thing about the lattice network is that with this particular configuration, any one terminal can be grounded without disturbing the balance

Fig. 3 shows a lattice network made up of transmission line sections. A shorted line is shown for the inductive arm, and an open line for the capacitive arms, but this is not the only configuration that can be used; it is merely the one which results in the least total amount of line. Its bandwidth characteristics are also fairly good. The reactance of the shorted line is jZ_oT and that of the open line is $-jZ_o(1/T_1)$, where T and T_1 are the tangents of the angles θ and θ_1, respectively. For the symmetrical lattice matching section it is required that $X = \sqrt{R_{in}R} = Z_oT$, and also that $TT_1 = 1$, from which it can be shown that the total line length $\theta + \theta_1$ must be 90°, or a quarter wavelength. The tap point is given by

$$\tan\theta = \frac{\sqrt{R_{in}R}}{Z_o} \quad \text{and} \quad \theta_1 = 90 - \theta$$

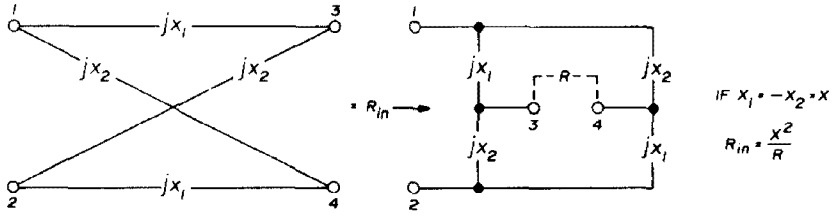


fig. 2. Reactive lattice network. In this configuration any one terminal can be grounded without disturbing the balance to ground. The transmission-line balun shown in fig. 3 is based on this equivalent circuit.

For the greatest bandwidth, tan θ must be small rather than large. That is, the tap point must be near the shorted end of the total line in each case. This dictates that Z_o must be as large as convenient. In the special case where $R_{in} = Z_o$, where the same line impedance is used for matching as in the transmission line to be matched, tan $\theta = \sqrt{R/Z_o}$.

The curves of fig. 4 give the electrical length of the tap point for some typical Z_o values and antenna resistances, assuming you want to match to 50-ohm coax. The *physical* length, of course, depends on the frequency and v_p, the velocity of propagation on the line (typically 0.82 for twin lead, 0.66 for coax, and 0.96 or so for open-wire feeders). In feet, it is 2.73 $(v_p/f)\theta$ where f is in MHz and θ is in degrees.

Fig. 5 shows one configuration as connected to and suspended from a balanced antenna. For a center-fed full-wave antenna of R = 4000 ohms and 300-ohm twin-lead for the balun, $\theta = 56.15°$. The tap point is thus at 62.4% of the total length of the quarter-wave twin lead, measured from the shorted end on each line.

re-entrant crossed line

The re-entrant crossed line is another form of transmission line impedance-matching balun that may be convenient in some cases, especially where you may want to feed a grounded vertical array using open-wire line. In this case the total length of line required turns out to be a full wavelength. The line and its equivalent circuit is shown in fig. 6; note the *crossed* connections.

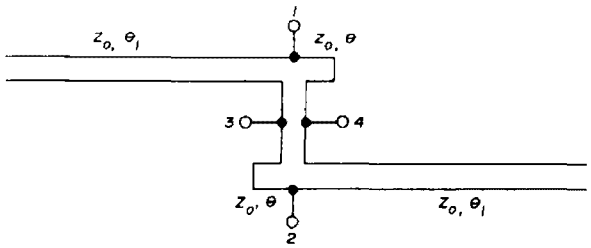


fig. 3. Open-wire transmission-line balun using the lattice network. Antenna is connected to terminals 1 and 2; transmission line is connected to terminals 3 and 4. Tap point for terminals 3 and 4 for matching 50-ohm coaxial cable is given in fig.. 4.

It turns out that for the case of interest, namely, a symmetrical lattice with arms equal and of opposite sign, the impedances across terminals 1-2 and 3-4 in the equivalent circuit of fig. 6 must be anti-resonant. That is, $T = -T_1$, where in this case the Ts are tangents of *half* the angle. You are left with the lattice itself in fig. 6, and for $S = -S_1$ simultaneously with $T = -T_1$ it is necessary that $\theta = 360 - \theta_1$. The result is

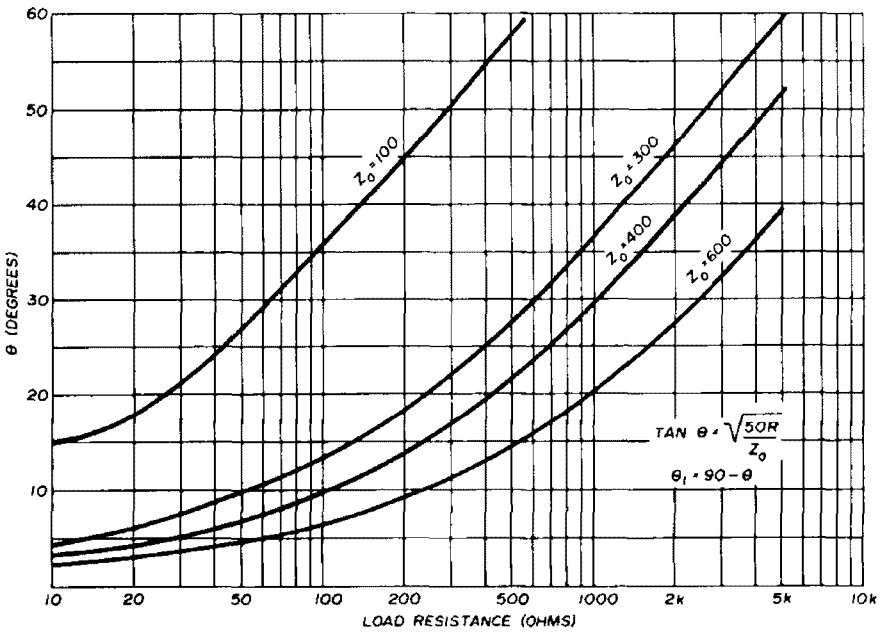


fig. 4. Balun tap point for matching 50-ohm coaxial cable to the open-wire balun shown in fig. 3.

$$\sin\theta = \frac{2\sqrt{R_{in}R}}{Z_o}$$

$$\theta_1 = 360 - \theta \text{ in degrees}$$

For this case Z_o must be greater than $\sqrt{R_{in}R}$ for a real result since sin θ cannot be larger than unity. This lattice can also be used as a balun — any one terminal of the network can be grounded without disturbing the balance to ground.

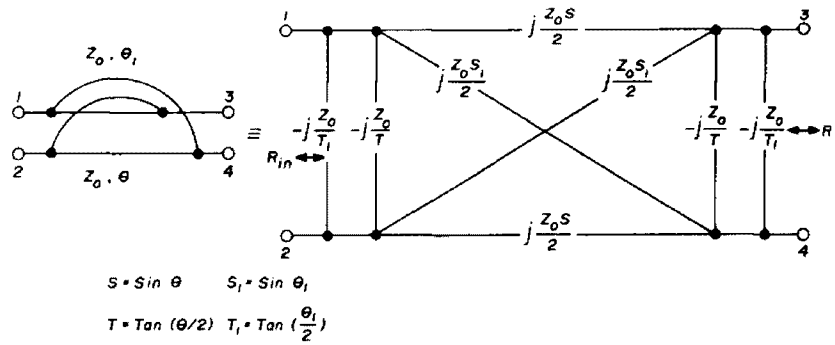


fig. 6. Crossed re-entrant line impedance-matching balun discussed in the text.

If it is assumed that you are matching to 50-ohm coaxial line, then for various characteristic impedance values for the re-entrant balun, you can match load resistances, R, up to the values shown below:

Z_o	R_{max}
300	450
460	1058
600	1800
800	3200
1000	5000

The maximum bandwidth will be achieved when the lattice arm reactance variation with frequency is the least. Since each arm is proportional to sin θ, it follows that when the variation of sin θ is the least, broadest bandwidth will be attained. This occurs when sin θ is near unity, which means you ought to choose a value for Z_o somewhere near (but always greater than) $2\sqrt{R_{in}R}$.

For $Z_o = R_{in}$, sin $\theta = 2\sqrt{R/Z_o}$ and Zo can be no smaller than 4R. This matching balun is not too well suited, therefore, to very high impedance antennas.

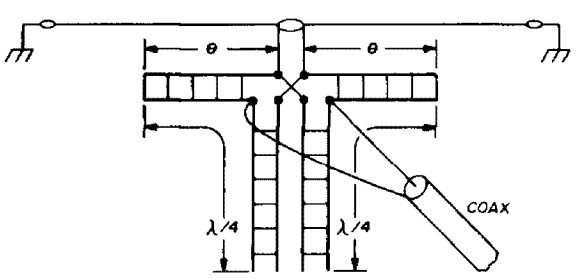


fig. 5. Typical open-wire impedance-matching balun based on the simplified circuit in fig. 3. In this case the balun is suspended from the center of a balanced antenna.

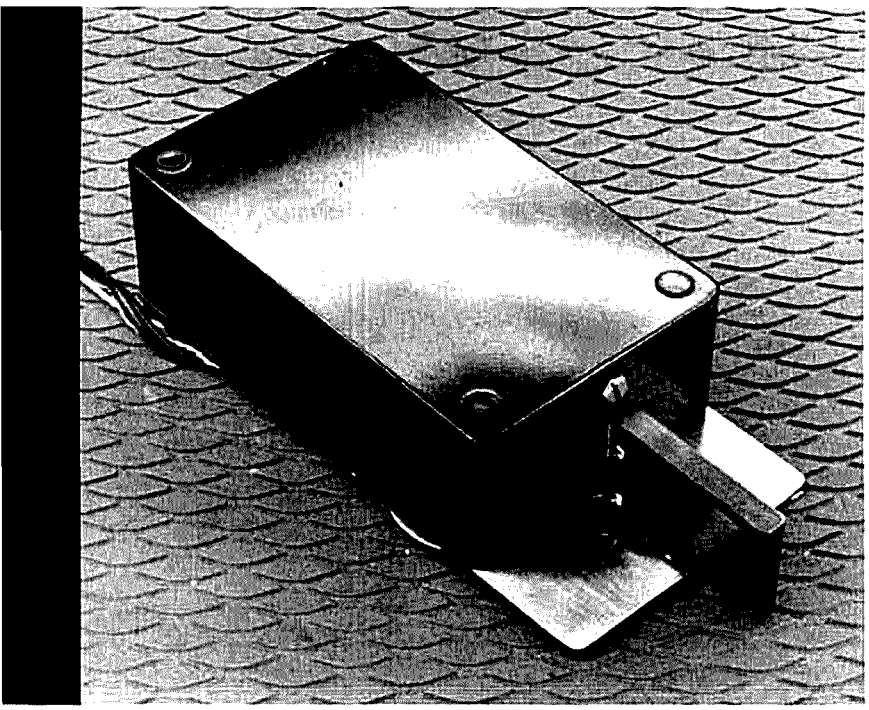
For an antenna resistance of, say, 400 ohms, to match a 50-ohm coax using 300-ohm twin lead for the balun requires θ = 70.5° and θ_1 = 289.5°. The equivalent lattice arms are (300 sin 70.5°)/2 = 141.4 ohms and (300 sin 289.5°)/2 = - 141.4 ohms. Transformation is $(141.4^2)/400$ = 50 ohms. The reactances across the input and output terminals are (300)/(tan 70.5/2) = 424.3 ohms and (300)/(tan 289.5/2) = - 424.3 ohms, and are in anti-resonance.

As a final example, suppose you wanted to feed a grounded vertical antenna having a base resistance of 36 ohms, using 300-ohm twin-lead, and also wanted to use twin-lead for the balun matching section. θ = 43.85° and θ_1 = 316.15° for this case, so for 7.2-MHz operation you would need 13.64- and 98.32-feet of twin-lead for the two lines (remember to cross the connection).

references

1. R. W. Johnson, W6MUR, "Two-Band Antenna Matching with Stubs," *ham radio,* October, 1973, page 18.
2. R. W. Johnson, W6MUR, "Transmission Line Matching Using Two Fixed Capacitors," *ham radio*, September, 1973, page 58.

ham radio



compact electronic keyer package

Complete electronic keyer, ac power supply and squeeze paddle in one compact package

Gene Brizendine, W4ATE, Huntsville, Alabama 35803

As the panarama of amateur equipment design flows on, transceivers and compact solid-state power supplies have rendered operation from your boat, car or camper increasingly attractive. Strangely enough, in this enlightened age of integrated circuits, the electronic keyer has hardly kept pace, and it generally comes in two packages, or one bulky one.

Early labor-saving electronic keyers required sizeable power supplies and heat-radiating tubes and required considerable operating desk space. The keyer described in this article emphasizes compact packaging, rather than electronic features. Briefly, what is probably the most advanced squeeze paddle circuit, including 16 ICs, power supply and monitor, is housed in a package the size of a conventional bug. The circuit is essentially that used in the Pickering keyer.[1]

construction

Printed-circuit board is used extensively because of its rigidity, shielding and

ease of working. A sheet of PC board forms the base, which supports all components. An inverted bakelite instrument case, approximately 6x3x2 inches, forms the cover. The cover is shielded with a sheet of heavy aluminum kitchen foil, cemented to all inside surfaces and extending over the edges, where it is clamped against the base in final assembly.

The RTL logic was wired in when the keyer was designed several years ago. However, an even more compact design is now possible with DIP devices and printed-circuit techniques.

To simplify the IC wiring, perforated board is used, with holes punched in a 45-degree pattern. Eight leads of each IC are inserted into holes forming a square, with pin 8 (+) at the topmost corner and the ground lead, pin 4 (−) at the bottom corner. Thus, IC power wiring is completed by joining all top corners with a positive voltage bus and connecting the bottom corners to ground. This method quickly connects 32 leads and also allows easy identification of the remaining pins.

The 3-volt IC power supply components are mounted on vertical strips of PC material which are soldered to the base. A pocket transistor radio output transformer serves as a compact power transformer. After completing the IC board and power supply wiring the board is secured to the base with two studs.

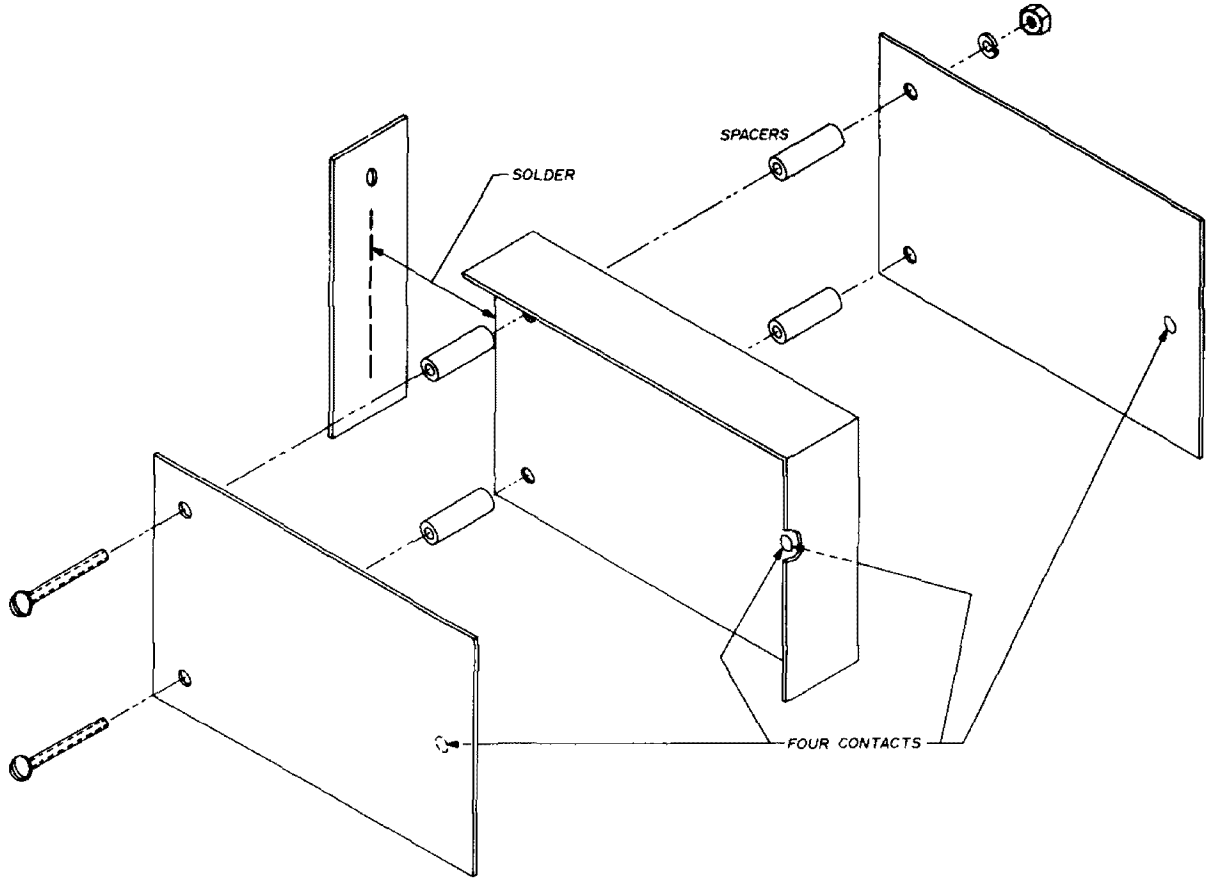
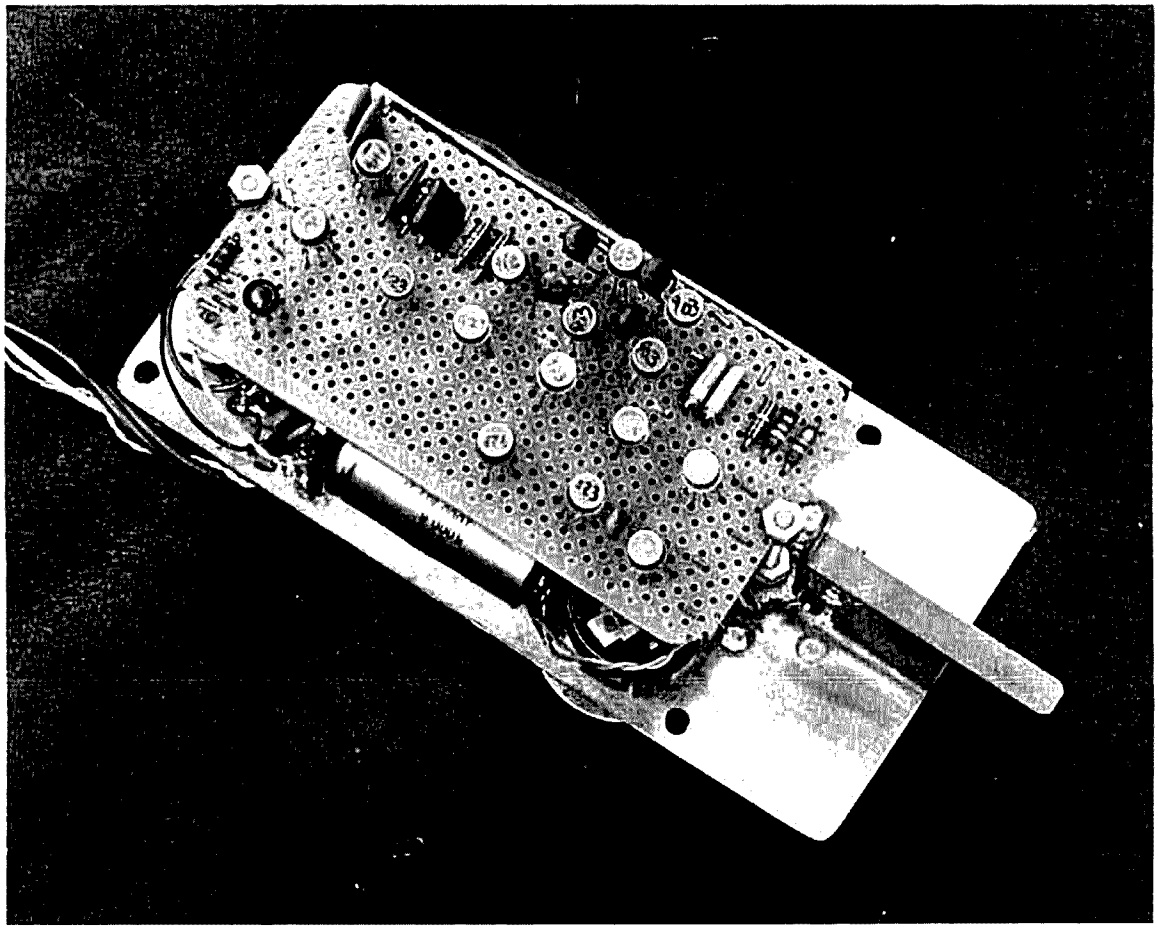


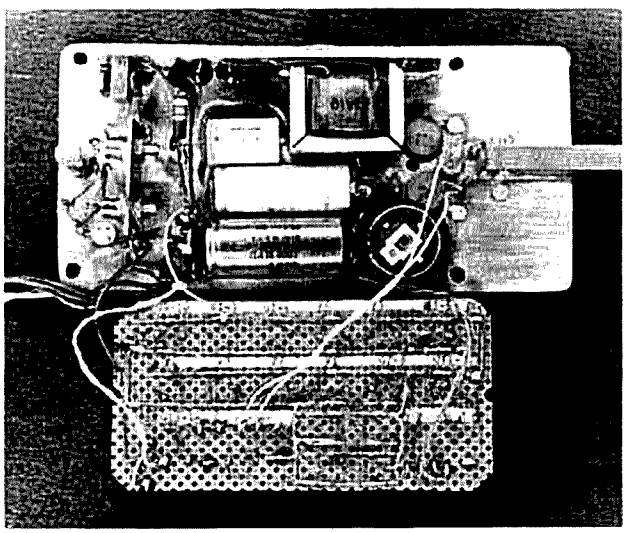
fig. 1. Squeeze paddle details. The middle double-sided PC board carries the stationary contacts. Single-sided PC board is used for the dot and dash paddles.



Keyer with the cover removed showing the ICs and discrete components mounted on perforated circuit board. The speed control is located at the bottom left side of the keyer (foreground).

squeeze paddle

The squeeze paddle assembly is also constructed of PC board as shown in fig. 1. The double-sided stationary portion is made very rigid by soldering strips along the front, top and rear edges. The bottom edge and the rear strip are soldered to the base. The rear strip is also secured to the cover with one screw.



Interior of the electronic keyer showing the arrangement of the power-supply components and the speed control.

Coin-silver contacts are soldered at the front edge of the boards to provide maximum travel when pressure is applied at mid-paddle. Some experimenting with spacers between paddles may be necessary to satisfy your particular preference. If audible clicking is objectionable, the inner surfaces may be covered with sound-absorbing felt or flocking spray.

The light, compact package shown may be attached on the operating desk with double-sided masking tape. At my station it has given four years of carefree operating pleasure.

reference

1. T. Pickering, W1CFW, "The Micro-Ultimatic," *73*, June, 1966, page 6.

ham radio

calculating gain vs height of DX antennas

Using a simple graph to predict the gain of your DX antenna if you raise its height above the ground

Gerd H. Schrick, WB8IFM, 4741 Harlou Drive, Dayton, Ohio 45432

It is well known that the higher your antenna is, the better it is for working DX. Although some amateurs talk about optimum antenna height, it is generally accepted that the higher the better. The question is, how much better?

Recently, I replaced an 80-foot tower with a 100-foot tower; both towers were available temporarily so I could make some comparison measurements. With the same antennas mounted on both towers, any gain differences had to be attributed to the difference in height. A gain difference of 2 dB was consistently measured, whether I was working local stations or DX. However, comparison measurements with DX stations were more difficult because of fading and other long-distance communications problems.

theory

Any horizontal antenna works in conjunction with the ground and the mirror-image concept can be applied. The formula for radiated power density, P_d vs the radiation angle, a, is

$$P_d = \sin^2 (h^\circ \sin a) \qquad (1)$$

where h° is antenna height in degrees of wavelength (1 wavelength = 360 degrees).

The gain difference, in dB, for two different heights, h_1 and h_2, is:

$$\triangle dB = 10 \log \frac{P_{d2}}{P_{d1}} = 20 \log \frac{\sin (h_2 \sin a)}{\sin (h_1 \sin a)}$$

For DX work a very small angle of radiation, a, a so-called grazing angle close to zero degree, can be used. Therefore, sin a is a very small number, much less than 1. The factor $\sin^2$ becomes very small when the actual antenna height is in the order of a few wavelengths. This results in the following approximation:

$$dB_{DX} = 10 \log \frac{h_2^2}{h_1} = 20 \log \frac{h_2}{h_1} \quad (3)$$

This is a surprisingly simple formula. Since the wavelength factor has dropped out, it is no longer necessary to express antenna height in electrical degrees — absolute feet or meters will do. Doubling your present antenna height will give you a maximum gain of one S-unit (6 dB). This is shown in fig. **1.**

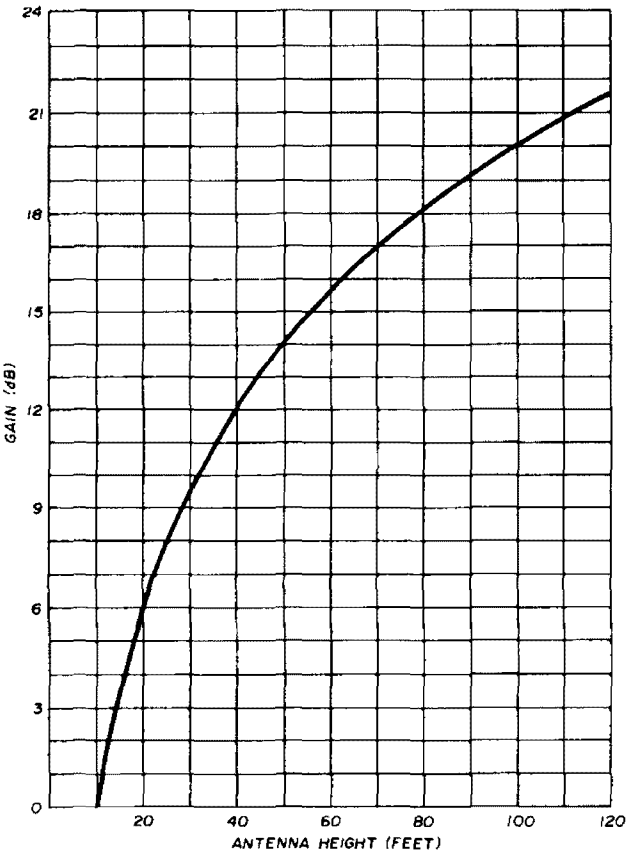


fig. 1. Antenna gain in dB for DX obtained by increasing height of an antenna above ground (angle of radiation less than 3 degrees). Referenced to zero dB gain at 10-feet antenna height.

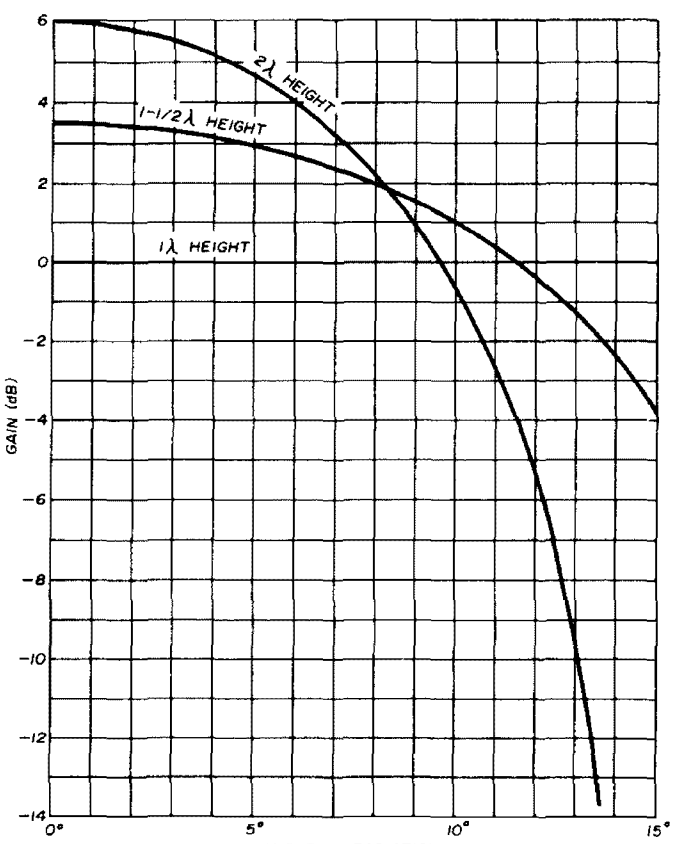


fig. 2. Power density vs angle of radiation for antennas 1½- and 2-wavelengths above ground, as compared to an antenna 1-wavelength above ground.

example

In my case, where I increased my tower height from 80 to 100 feet, the expected power gain is predicted by **eq. 3:**

$$dB = 20 \log \frac{100}{80} = 20 \log 1.25 = 1.9 \text{ dB}$$

This is in very good agreement with the practical measurements described previously.

Fig. 2 shows the difference in power density in dB versus the angle of radiation for horizontal antennas 1-, 1½-, and 2-wavelengths above ground. This graph is based on the use of the exact formula given in **eq. 2.** It can be seen that the approximation in **fig. 1** is quite accurate for radiation angles less than about 3 degrees. Certainly, it is not possible to obtain higher gain than that implied by **fig. 2.**

ham radio

antenna and control-link calculations for repeater licensing

How to calculate the antenna patterns and control links required for your new FCC repeater license

P. J. Ferrell, W7PUG, 6021 South 119th Street, Seattle, Washington 98178

The new FCC repeater regulations are admittedly a tough act to follow. In particular, the antenna pattern and control-link power justification seem to be difficult to obtain. These items may be calculated directly rather than measured, since either method is acceptable to the FCC. No attempt will be made to derive or justify the formulas employed, but references for each are included wherein they are most adequately treated. Without attempting to ruffle any feathers, I feel that anything tending to raise the amateur's technological level above that of appliance operation and CB can't be all bad.

the antenna pattern

The first problem is to produce an "antenna pattern" and the supporting data to satisfy 97.41 (f)(3,6 and 7) of the new repeater regulations. Any repeater group possessing an antenna range or instrumented helicopter can pass immediately to control-link calculations on page 60. Antenna patterns can be measured of course, but great pains must be taken to guard against reflections, to be certain that only the far-zone pattern is sensed and that the instrumentation is in calibration. All in all, it would require an effort far beyond the meager resources of our group.

A calculated pattern approaches the actual radiation pattern only when the antenna is well clear of obstructions. This condition tends to be met only at vhf and above. Calculating the radiation pattern for an arbitrary current distribution is an extremely messy proposition, but if the antenna can be characterized as a linear array with equal current elements, considerable simplification can be obtained using pattern multiplication. Simply stated, you can separately calculate the patterns due to length, to width and to the individual elements and obtain the overall pattern by multiplying them all together,

provided the elements and their currents are all the same. Fortunately, the two most popular types of base-station antennas favored for repeater use meet this requirement.

Type 1, the J-pole or vari-loop antenna consists of dipole type elements mounted on a metal pole, separated about one wavelength and fed with a corporate feed insuring a nearly uniform and inphase current distribution.

Type 2 is typified by the Communications Products "Stationmaster" and the Prodelin "Big Stick" antennas. Homebuilt antennas of this type have been described in the amateur literature.[1] Some amateurs have "improved" them by specifying foam-dielectric coaxial cable, thereby insuring a non-uniform current distribution and a fair sized reduction in antenna performance. These antennas consist of 8 or 9 pieces of alternating series connected coax, each an exact half-wavelength internal electrical length. Special end elements are on the ends, and all elements are spaced about 0.3 wavelength apart and can be considered to have a shortened dipole pattern.

The vertical (E-plane) pattern of the antennas consists of an element part and an array part. We shall assume a *cosine* voltage pattern for the individual element. This is exact for a short dipole, and is a good approximation to a half-wave dipole. The normalized array part is given by:[2]

$$E = \frac{\sin(N\psi/2)}{N \sin(\psi/2)}$$

where $\psi = 2\pi D_e \cos \emptyset$,
N = number of colinear elements
D_e = interelement spacing in wavelengths
$\emptyset$ = elevation angle measured from broadside

The overall normalized E-plane voltage pattern is

$$E = \frac{(\cos \emptyset) \sin(N\pi D_e \cos \emptyset)}{N \sin(\pi D_e \cos \emptyset)}$$

You can see that this pattern depends only on the values of N (number of elements) and D_e (spacing between any two elements).

The horizontal (H-plane) pattern of antenna type 2 is omnidirectional and that of antenna type 1 can be adjusted to be omnidirectional. It can also be adjusted to give either a figure-8 or a cardioid pattern. These last two cases are difficult to handle because of the reflections occurring due to the support pole.

A fair approximation may be obtained by replacing the reflecting pole with an *image* antenna fed with an equal current, leading in phase by twice the spacing between the element and the support pole. A typical value for element spacing is about 4 inches or 0.1 wavelength. The normalized voltage pattern in this case is given by:[3]

$$E = \cos [2 \pi D_s(\cos \theta - 1)]$$

where D_s = element spacing from the pole in wavelengths
θ = azimuth angle

Note that if we set $D_s = 0$, then $E = 1$ for all azimuth angles, which is just the omnidirectional case.

Having obtained both E- and H-plane normalized voltage patterns, all that remains is to calculate the gain associated with each and multiply them (add dB) to obtain the overall gain.

Gain is calculated by adding up the total power (integrating) at all angles, recalling that power is proportional to voltage squared. Gain is defined as the ratio of maximum to average (isotropic) power. The FCC regulations specify a half-wave reference dipole rather than the more common mythical isotropic antenna, so a factor of 1.64 (2.2 dB) is tossed in to meet that requirement. This is due to a half-wave dipole having a gain of 2.2 dB over an isotropic antenna.

Even this greatly simplified calculation is likely to prove tedious, and so the job was subcontracted to a digital computer. The *Tymshare Superfortran* program used is listed along with sample printouts in the appendix. The two examples are for a 4-element J-pole set for a cardioid pattern, and a 9-element (Stationmaster)

antenna. Both E- and H-plane patterns are symmetric about the point of maximum gain (0°) so that only half of each need be printed out.

A polar plot of antenna gain in dB, or field voltage for each pattern can be constructed from the printout. If no reflecting mount is present, answer zero when asked "offset from reflecting mount, wavelengths." In this case, no H-plane pattern is printed out. However, it would be wise to construct a circular plot entitled "azimuth pattern" for submission along with the E-plane pattern.

the control link

I now wish to demonstrate compliance with 97.67(b) of the FCC regulations for a repeater control link. A calculation of this sort is commonly referred to as a "power budget." The handiest procedure is to use decibels exclusively in the magic formula:[4]

$$A = 36.6 + 20\log F + 20\log D + M + L_t - G_t + L_r - G_r$$

where A = worst case loss in dB between transmitter and receiver
F = frequency in MHz
D = separation between antennas in statute miles
M = fading margin in dB (typically 20)
L = transmission line losses in dB
G = antenna gains above an isotropic reference in dB

Clearly, antennas, transmission lines and the propagation path can be all be represented together as an attenuator of A dB.

A certain calculable amount of signal power is required at the receiver input to insure satisfactory link performance, and the necessary transmitter power output is just A dB above that amount. The required receiver input signal power may be calculated from:

$$R = -204 + 10\log B + NF + CNR$$

where R = required input signal power in dBW (dB below 1 watt)
B = receiver i-f bandwidth in Hz
NF = receiver noise figure in dB
CNR = required i-f carrier-to-noise ratio in dB to obtain satisfactory receiver performance. A typical value for an fm receiver is 12 dB

Finally, the required transmitter power is

T = R + A, T in dB relative to 1 watt (dBW)

The following numerical example was submitted to the FCC with the Seattle Repeater Group's application for a repeater license. At 450 MHz, for the given 25-mile path, loss between isotropic antennas is given by 36.6 + 20 Log F + 20 log D.

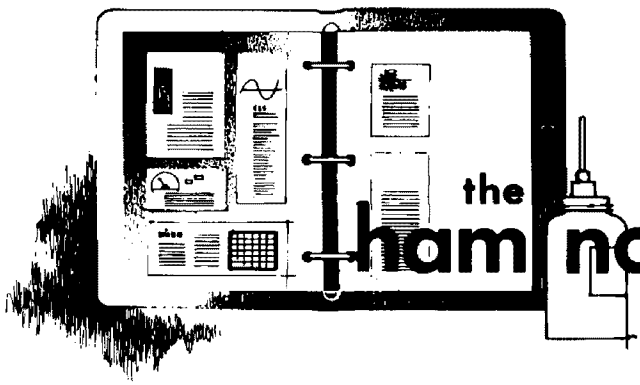
118 dB	median path loss, Green Mountain to Seattle
2 dB	repeater uhf receiving antenna gain (dipole)
4 dB	repeater site transmission line loss
9 dB	control site antenna gain
5 dB	control site transmission line loss
116 dB	median loss between units
20 dB	fading and diffraction loss margin
136 dB	worst case path loss between transmitter and receiver
-159 dBW	thermal noise power in 32 kHz i-f bandwidth (KT = -204 dBW/Hz)
15 dB	receiver noise figure
15 dB	required SNR at detector input for reliable tone transmission
-129 dBW	signal power required at the uhf control receiver

The necessary control transmitter power is 136 - 129 = +7 dBW. Converting to watts, 7 dBW = 5 watts.

references

1. K. W. Sessions, K6MVH, "Colinear Gain Antenna for VHF/UHF Repeaters," *73*, July, 1971, page 42.
2. John D. Kraus, *Antennas*, McGraw-Hill, New York, 1961, page 78, equation 4-52.
3. op. cit., page 292, equation 11-55.
4. *Reference Data for Radio Engineers*, Howard W. Sams & Co., Inc., Indianapolis, 1968, page 26-19.

ham radio



the ham notebook

RTTY line length indicator

When using radio teletype it is my practice to try to type some of my reply while printing the transmission of my contact. To do this I use a model 15 typing unit which prints the incoming signal while I am using the keyboard to punch a tape on my model 14 reperforator. As I cannot see what I am typing or tell when I reach the end of a line with this system, I have designed a system to count the words or spaces in the usual teletype line.

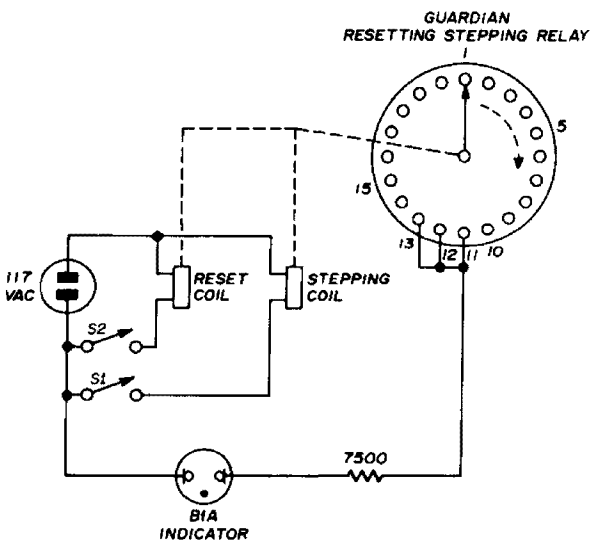


fig. 1. Circuit for the RTTY line length indicator uses a Guardian resetting-type stepping relay. Microswitch S1 is mounted under the space bar. Microswitch S2 is located under the carriage-return key.

The key to counting the number of words in a line is a stepping relay. I use a Guardian resetting type relay. Two microswitches are used to actuate the relay coils. These are mounted under the keyboard of the typing unit and are so positioned that when the space bar is pressed it also strikes the microswitch lever. This switch then closes the circuit to the advancing coil on the stepping relay, advancing it one step.

The other microswitch is positioned under the carriage return key. When this key is pressed this switch actuates the circuit to the relay reset coil. The relay is then reset and is again ready to count the spaces in the next line.

The usual practice is to have ten words per line. I connected my stepping relay to make contact on the 10th, 11th, 12th and 13th steps. When this contact is reached, the alarm circuit is energized. A bell, buzzer, light or other signal can be used. As this relay is a 115-volt ac unit, I use a GE B1A neon bulb in a red panel mounting. This bulb must be used in series with a 7500-ohm resistor. The light is mounted next to the keys and the frame of the typing unit.

After ten spaces or steps the light comes on and stays on for the next three steps. If advanced any further it will go out and you will know you have gone beyond the usual line length. You don't have to worry about over advancing the relay. It goes to the limit and that's all. The stepping relay should be mounted in an insulated box for quiet operation. I also shock mounted mine.

Hal Dressel, W2UVF

ssb filters

Recently, a short item on sideband location was published in *ham radio*, based upon information published in the instruction manuals for the Swan 500 and the Collin S-Line. There was mention of the need to maintain an adequate separation from a band edge. There are several other aspects of filter performance, some of which are easily demonstrated in a receiver or transceiver. By making these tests, a phone operator may have a better understanding of just where his sideband lies, and where he may produce or receive interference.

Without an antenna, or on a dead band, turn on the calibrator and set the dial to zero-beat with the signal. Then, turn the marker to that setting. Now, swing the dial and note the S-meter reading for the maximum signal strength off to one side. This maximum may be adjustable by the preselector tuning control, so that the bandwidth can be noted for different signal levels.

Turn the dial away from zero-beat until the beatnote has dropped to one of several convenient levels, and finally disappears in the noise. Make a note of these dial settings. Then change to the other sideband and repeat the test.

In the case of my Collins S-Line, the normal 2.1-kHz filter (for receiving only, it has also the 1.5 and 3.1 kHz filters) drops down or out about as follows:

Peak signal	LSB	USB
S7	+3.5 kHz	-3.5 kHz
S9 + 10 dB	+4.2 kHz	-4.2 kHz
S9 + 20 dB	+4.4 kHz	-4.4 kHz
S9 + 20 to S3	+3 kHz	-3 kHz

The size of an S-unit must not be taken as 6 dB. It is about 4½-dB on the Collins 75S3A meter at S9, and less than 3dB around S3. The marked range of 20 dB for S3 to S9 is about correct on my meter, indicating an average of 3.3-dB per S-unit. It takes only 10 dB to increase the meter from 40 to 60 dB as marked on the scale, but 20 dB above that.

The amount of interference, produced or received, varies with the power and the conditions. If a station talks locally or under good skip conditions with an S9+20 dB signal, he should reduce power as soon as he hears that he is above S9, or suspects it from the strength of the station being worked. Minimum power is required by the Communication Act, Section 324. Otherwise, there may be considerable unnecessary "splatter" due to the filter skirts or equipment problems. In addition, there may be off-frequency emissions greater than those indicated by the filter skirts demonstrated by the above tests. Furthermore, the transmitter can have a somewhat different filter performance than the receiver.

FCC regulations, Section 97.63, says ". . . sideband frequencies . . . shall be confined within the authorized amateur band." It doesn't specify any number of decibels down. This means that the amount of attenuation that is acceptable, depends upon power and propagation conditions, not just -20 dB or -60dB, or any other particular figure. Some standards have been set — BBC Johore, in discussing their harmonic on 14240 kHz, indicated that ITU requires harmonics to be down 60 dB.

The test demonstrates that you must not operate as close to a band edge as previously suggested, even when you have the equipment to comply with the requirement that the frequency of the emissions shall be measured by independent means, and of an accuracy sufficient to assure operation within the band.

The test also shows that interference may be expected from other stations in the range given in the above table, depending upon his filter performance; and that you will produce interference on the *opposite* side by at least the amount shown, plus any additional hash produced above the filter curve. If you hear the calibrator on the "undesired" side of the zero beat, then you may expect to receive some interference there, and produce some on the opposite side when transmitting, because of the lack of pure *single* sideband operation.

Bill Conklin, K6KA



comments

frequency scaler

Dear HR:

Response to the construction article for the simple frequency scaler appearing in *ham radio* for September, 1972, has been most gratifying. That article stirred up the interest of a large number of hams and for many of them it was the impetus for their first-time experience with frequency counters as well as with scalers. To the serious experimenter the article presented no problems. Some difficulties cropped up for the less knowledgeable, however. One of these was that their high-sensitivity, high-impedance counter might indicate twice or even three times the frequency which was fed to the scaler.

Although the solution to this type of problem is really quite simple, a wide variety of schemes have been tried by amateurs in an attempt to solve this problem. One of these schemes is the one which W6MGI described in *ham notebook* of the February, 1973, issue (page 57). For his particular cable and for its particular length, it worked for him. His L-section approach does not provide a general solution, however. The characteristics of his filter will change with every foot of cable he adds or removes (50-ohm cable, for example, has about 28.5 pF per foot).

Any ECL or other very high-speed pulse circuit requires detailed attention to feedback and matching problems. Use of a ground plane which provides a good, low-impedance ground current return path is one essential. A well regulated power supply is another. These can be and are provided within the scaler itself.

The matching problem can best be met outside of the scaler. Experience with and during development of the scalers manufactured by Belmont Spectrum Research clearly demonstrates that for interconnects of more than about 5 inches (some of this length is within the scaler) the matching problem can only be generally successfully met by providing an adequate line termination. This, of course, cannot be accomplished within the scaler. It is therefore standard practice for manufacturers to caution users that an interconnecting cable (scaler to counter) must be terminated in its characteristic impedance if waveform distortion cannot be tolerated. Whether it is called distortion, ripple, ringing or reflections, is immaterial; the steep multiple wave fronts of ECL or other high-speed pulse systems are what trigger the counter and disconcert the casual scaler user. So — the really simple solution to the matching problem is to terminate the line.

One means of providing a proper termination is to connect a resistor, equal to the characteristic impedance of the line you are using, across the input connector inside the counter. Another means, which is preferred because its use is more flexible and does not entail any change to the counter, is to use a "termination adaptor" at the counter end of the line. This method will provide a proper termination when the scaler is used and when unplugged will permit retention of the counter's regular input for use over its normal range. A BNC tee-connector (UG-274/U) at the counter, with the appropriate resistor connected to one arm of the tee, will do very nicely.

These tees are often available through surplus sources at very low cost.

An even better arrangement is to use the 50-ohm termination adaptor made by Tektronix (part no. 001-0049-01). This adaptor, originally intended for use with Tektronix scopes, is highly recommended. It currently sells for $10.

Now, as to overall results. My Belmont Spectrum Research scaler will drive up to 18 feet of either 50- or 100-ohm coaxial cable without substantial waveform degradation when any one of these terminating arrangements are used.

F. Everett Emerson, W6PBC
Belmont Spectrum Research
Belmont, California

passive sideband generator

Dear HR:

Although I have been running my passive sideband generator at the signal levels indicated in the article,* (about 3 volts peak rf and 0.3 volt peak audio at the input ports), the audio level should be several volts. I forgot about the extra loss in the resistive output branch of the audio phase shift/network.

An appropriate audio signal level can be deduced as follows. Suppose the peak rf input is 3 volts. After going through the rf phase-shift networks this is reduced to 2.1 volts. A good guess at the modulator resistance is 1000 to 1500 ohms (500 for the balance pot, 200 or so for the diode and 300 to 800 for the load seen through the output rf transformer).

If we pick 1250 ohms, the rf current through a diode is about 1.7 mA. The modulating signal should be small compared to the carrier, ten percent being a usual limit, so the audio peaks should approach, but stay under, 0.17 mA.

The resistance of the audio phase-shift network output branch, including the modulator impedance, has been adjusted to 3900 ohms. Hence, the audio voltage at an output of the phase shift network should be about 0.66 volt. As noted in Van Heddegem's article, the voltage across the 3-ohm input resistor will then be 1.32 volt, making the total peak input voltage 6.4 volts (about 1.4 watts). This is an upper limit. A third to half of this voltage is adequate while still being ten times what I have been using.

Incidentally, with this higher level of modulation, the carrier balance will be much less critical since a voltage gain of ten or so is thus moved from the sideband amplifier to the audio section.

Worthie Doyle, W7CMJ
Port Orchard, Washington

Dear HR:

Cheers to author Doyle for his fine article on the phasing-type ssb generator in the April issue. Some three or four years ago I built up a solid-state phasing exciter (not exactly passive like Doyle's) and I used a tired 6L6 linear amplifier. I worked Albuquerque, New Mexico, from Libertyville, Illinois, and was heard in Guantanamo Bay, Cuba, on 14 MHz.

For checking the Doyle passive audio phase-shift network I recommend use of an oscilloscope having separate X and Y axis inputs. Leave the *carrier input* of Doyle's **fig. 1** disconnected for this test. Disable the sweep of the oscilloscope, and adjust both channels for equal gain. Hook the X input to one side of the phased audio output; the Y input to the opposite side. Talk into the microphone (or use a signal generator) and observe the pattern on the scope from 300 to 3,000 Hz. It should remain a nice circle, changing in diameter with signal level. If not, make adjustments to the circuit until you get the best circle.

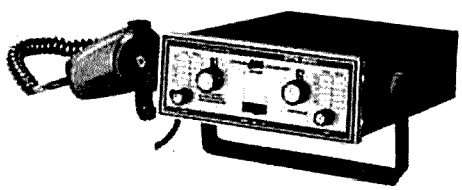
Connecting the phase-shift network ground terminal to the oscilloscope chassis ground should cause no change to the geometry. This same method may be used to check the ninety-degree rf phasing as well, but probe lead length and amplifier phase differences inside the oscilloscope can give misleading information.

Paul Schmidt, W9IDP
Libertyville, Illinois

*W. Doyle, W7CMJ, "Phasing-Type SSB Generator," *ham radio*, April, 1973, page 22.

new products

two-meter fm transceiver



A complete new line of American-made 2-meter fm transceivers for the amateur has been introduced by General Aviation Electronics, Inc. (GENAVE). The new offering includes the GTX-10, the GTX-200, and the previously introduced and very popular GTX-2.

The new GTX-10 fm transceiver is a full 10-channel, 10-watt output unit and retails for $199.95. The new GTX-200 features independent selection of 10 transmit and 10 receive frequencies, offers 30 watts nominal output power and retails for $259.95. The well-received GTX-2 provides 10 push-button channels with backlighting for night operation, 30 watts nominal output power and retails for $249.95.

The radios are manufactured in the same U.S. Government inspected facilities where precision aircraft instruments are fabricated, under the same watchful quality control procedures.

Internally, all radios are equipped with netting trimmers for each transmit crystal. All use standard, readily available, American-made semiconductors. High selectivity 8-pole second i-f filters are incorporated in the design of all three units, and rf output stages are vswr protected.

The GTX line is engineered for use with available tone encoders and autopatch service, as well as simultaneous operation on MARS frequencies. Externally, all three radios have multi-position switches which include a low power (one-watt) setting for longtime low-power drain operation, and indexed volume and squelch controls.

Each GTX transceiver comes complete with a quick-disconnect power cable, SO-239 antenna connector, mobile mounting bracket and sturdy ceramic plug-in microphone. A 146.94-MHz communications channel is also included. The remaining plug-in crystals are available at $6.50 each for installation at the factory or by the owner.

The new GTX-10 features superlative cross-mod performance, and is easily cross-wired for duplex crystal operation. The circuit board is laid out so that conversion to 30-watt output can be accomplished quite easily.

The new GTX-200, with independent selection of 10 transmit and 10 receive frequencies, offers 100 possible channel combinations. A switch for lock-in of preselected frequency pairs allows simple one-knob operation when desired. High sensitivity is assured by incorporation of a dual-gate mosfet in the receiver front-end. An external speaker jack is provided on the rear panel.

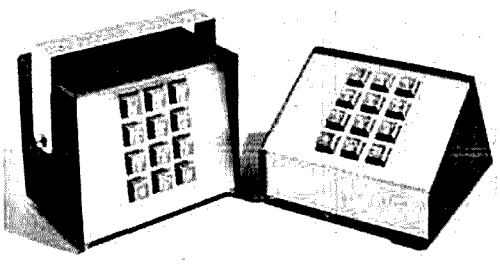
The popular GTX-2 has been refined and updated for superior sensitivity with the same dual-gate mosfet in the receiver frontend as is found in the GTX-200.

The new radios are founded on the technology and know-how derived from Genave's experience as a leading manufacturer of a full line of navigation and

More Details? CHECK–OFF Page 110

communications radiotelephones for the marine industry. For more information, write to General Aviation Electronics, Inc., 4141 Kingman Drive, Indianapolis, Indiana 46227, or use *check-off* on page 110.

touch-tone encoder enclosures

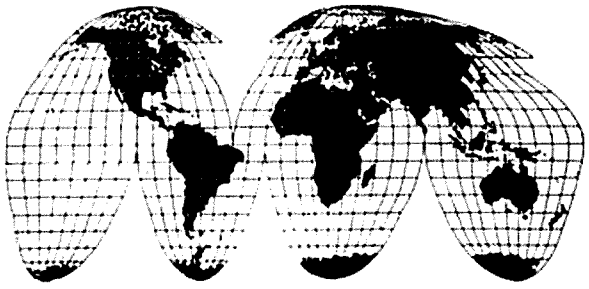


There has been a tremendous increase in the use of Touch-Tone encoders for use in autopatch and other special control systems in fm repeaters. Now two compact enclosures are available for Touch-Tone pads — one for base stations and one for mobile use. The Touch-Tone pad is held properly in position by pre-mounted internal brackets.

The top and sides of the enclosure are covered with walnut-grained vinyl. The satin anodized aluminum face is die punched to accept standard 12-button Touch-Tone pads such as those manufactured by Western Electric, Stromberg-Carlson, ITT-Kellogg, etc. (Automatic Electric Touch-Tone pads will not fit because they use non-standard spacing between the buttons.)

The mobile mount "M" Touch-Tone enclosure has an anodized pivoting gimbal bracket which provides multi-position mounting under the instrument panel, as well as top-of-equipment mounting on four rubber feet. The base station "B" enclosure holds the Touch-Tone pad at a convenient 30-degree angle. There is ample room inside both enclosures for transmitter keying circuitry. Either model is available for $6.25 including postage from the Detroit Area Repeater Team, Post Office Box 201, Clawson, Michigan 48017. For more information, use *check-off* on page 110.

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More Details? CHECK—OFF Page 110

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6" x 6", $1.10 ea. - 6 x 12, $2.00 ea. - 12 x 12, $3.75 ea.
New P.C. Boards — G10, Fiber glass punch:
F Pattern 4.5 x 6.5, .062 holes, 5 per 1" $1.30
P Pattern 4.5 x 6.5, .042 holes, 10 per 1" $1.35
G Pattern, 4.5 x 6.5, .062 holes, $1.30
Pkg. 10 Bircher P.C. Board, metal 2" slides $1.00
Package of 50 flea clips for above punched Boards, .062 ..... 75¢
30 ¼ or ½ W resistors, packaged 5 per value your choice of values $1.00
25-1W resistors, packaged 5 per value, your choice of values $1.00
15-2W resistors, packaged 5 per value, your choice of values $1.00
5 ¼ or ½ W, 1% resistors, packaged 5 per value, your choice of values ..... .50
5 ceramic disk caps, .001-01, packaged 5 per value, your choice of values ..... .50
5 mica dip caps, 1 pf-150 pf, packaged 5 per value, your choice of values ..... .50
5 mica dip caps, 180 pf-820 pf, packaged 5 per value, your choice of values ..... .75
5 mica dip caps, 910 pf-1500 pf, packaged 5 per value, your choice of values $1.00
Wire Kit #22 solid PVC, 6 spools, 6 colors, 50' ea. spool $3.50
Wire Kit # 22 stranded PVC, 6 spools, 6 colors, 50' ea. spool $3.50
Wire Kit #24 Solid PVC, 6 spools, 6 colors, 50' ea. spool $3.50
Wire Kit #24 stranded PVC, 6 spools, 6 colors, 50' ea. spool $3.50
10' — Ten cond. ribbon wire, color coded, #22 or #24, stranded $1.50
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C & K #7103 mini switch, SPDT on-off-on $1.20
C & K #7201 mini switch, DPDT on-on $1.35
C & K #7203 mini switch, DPDT on-off-on $1.55
Alco 105D MST momentary on-off-momentary on $1.25
Central Lab DPDT push momentary. SPEC. 4/$1.00
Connectors, PL259, $.45; PL258, $.70; 175U or 176U, $.20 ea.; UG 88 cu., $.50; UG 201 a/u (N to BNC adapter), $.75; RCA to UHF, $.90.
Encapsulated chokes 1uh to 5 Mh, choice 3/$1.00
Varo type mini bridge rectifiers, approx. ⅝" sq. size: 2 amp. - 50 v., $1.25; 4 amp. - 50 v., $1.25; 6 amp. - 50 v., $1.25; 2 amp. - 100 v., $1.25; 4 amp. - 100 v., $1.25; 6 amp. - 100 v., $1.25; 2 amp. - 200 v., $1.50; 4 amp. - 200 v., $1.50; 6 amp. - 200 v., $1.50; 2 amp. - 400 v., $1.50; 4 amp. - 400 v., $1.50; 6 amp. - 400 v., $1.50
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To-5 case, 1 amp. - 200 v., $.70 ea.; 1 amp. - 400 v., $1.00 ea.
SCR 200 v. - 8 amp. thermo tab ..... $.80 ea.

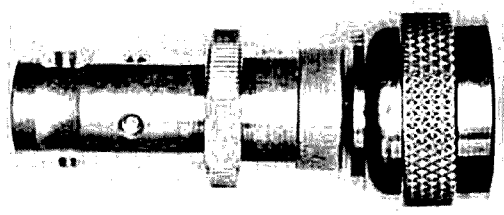
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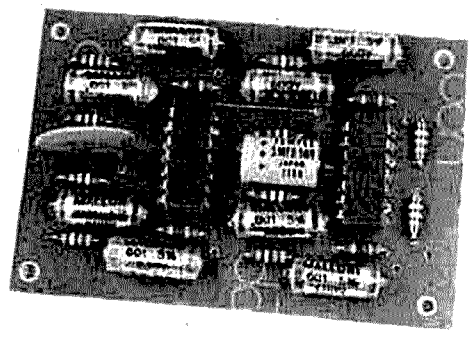
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connector adapters



A new series of adapters is now available which will interconnect any combination of cables and connectors having BNC, RCA phono, microphone (Amphenol 5/8-27 type) or type-F (CATV type) terminations. These adapters are intended for use in test equipment, and in audio, video and rf applications. The adapters are priced at $4.95 each, postpaid in the United States. Order from Adapters Unlimited, Post Office Box 48822, Los Angeles, California 90048. For more information, use *check-off* on page 110.

cw filter



The new audio CW filter available from MFJ Enterprises offers three degrees of switch-selectable selectivity in the same filter. The three bandwidths are 180, 110 and 80 Hz. In the 80-Hz position, response is down 60 dB one octave away (one-half and twice the center frequency of 750 Hz).

There is no insertion loss when this audio filter is switched into the line. Also, with most narrow-band audio filters ringing can make copying impossible. In the new CWF-2 CW filter ringing is nearly

More Details? CHECK–OFF Page 110

eliminated by the technique of cascading four low-Q stages. This provides very narrow bandwidth and extremely high skirt rejection without audible ringing.

The CWF-2 filter offers very low output impedance and a very high input impedance. This means that, unlike some other filters, no impedance matching is required for optimum performance. Loads greater than 500 ohms produce no distortion and loads less than 500 ohms, such as an 8-ohm speaker, produce some distortion which does not affect copying.s,

To use the CWF-2 you simply plug it into the phone jack of your receiver or connect it to the speaker terminals. It can also be installed between audio stages in your receiver. The circuit consists of four IC operation amplifiers in an active-filter design which eliminates all inductors and reduces the unit's size to a mere 2x3-inch printed-circuit board.

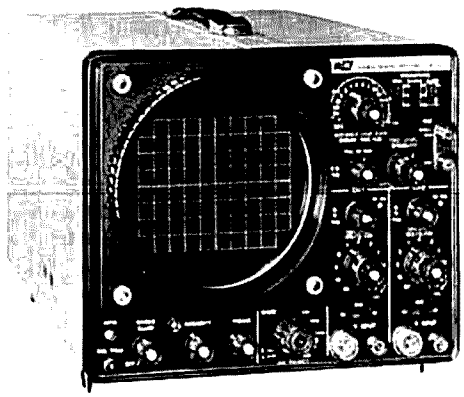
The CWF-2 CW audio filter is available in kit form for $9.95, or completely wired, tested and guaranteed for $12.95. For more information, write to MFJ Enterprises, Post Box 494, Mississippi State, Mississippi 39762, or use *check-off* on page 110.

two-meter collinear mobile antennas

Extremely low radiation angle, 5.2-dB gain over a 1/4-wave ground plane, low swr and wide bandwidth are features of Hustler CG-Series Super Gain two-meter collinear mobile antennas from New-Tronics Corporation.

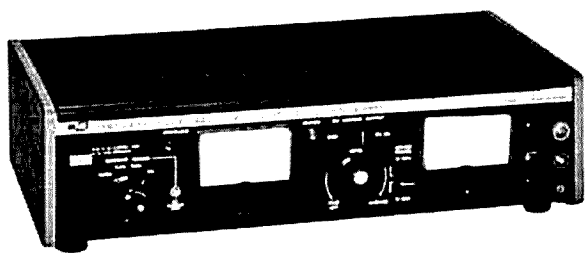
Model CGT-144 is complete system including collinear antenna with stainless steel radiating sections, 180° swivel ball, heavy duty trunk lip mount for easy "no holes" installation on side or edge of trunk lip, and 17-foot RG-58/U coax with factory-attached connectors. Power rated at 200 watts fm, the completely operational CGT-144 has a swr of 1.1:1 (typical) at resonance and a swr within 1.5:1 over its 6-MHz bandwidth of 143-149 MHz. Overall length is 86 inches.

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7406	.55	7450	.29	74145	1.25
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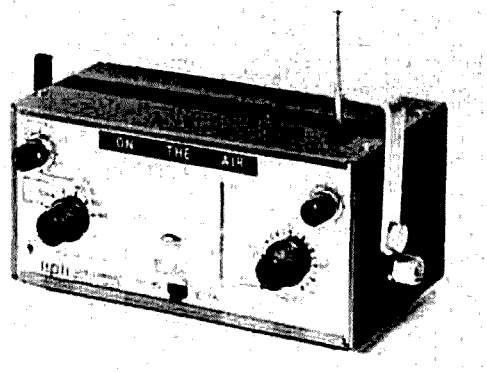
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The Hustler Model CG-144 consists of the 84-inch collinear antenna with 3/8-24 threaded base to fit standard mobile ball mounts. It has the same electrical characteristics as the Model CGT-144. For complete specifications contact New-Tronics Corporation, 15800 Commerce Park Drive, Brook Park, Ohio 44142, or use *check-off* on page 110.

base command



The PACE Communications division of Pathcom, Inc. today announced a new special purpose instrument called *Base Command* designed for sophisticated control of base station operation on the amateur two-meter band. Designed specifically to keep constant surveillance on the performance of your base station, the P5407 *Base Command* is placed in the transmission line between your transceiver and antenna to monitor and control the functions of your transmitter.

Antenna installation efficiency is measured by checking the standing-wave ratio. The transmitter power is measured on one of three scales; the 5-watt and 50-watt levels terminate in an internal dummy load. The 500-watt scale samples power while it's going through the line to the antenna.

The modulation capability of the transmitter is measured on the reference meter. The audio quality of your transmitted signal may be monitored continuously. The unit also provides a visual "on the air" indication. Performance of your hand-held or mobile sets can also be checked with the built-in field strength meter. A television interference filter is also built into the P5407.

More Details? CHECK–OFF Page 110

This is where the name, *Base Command,* was designated since this versatile instrument not only monitors but controls the base-station transmitter. For more details, write to PACE Communications, Box 306, Harbor City, California 90710 or use *check-off* on page 110.

phase-locked loop handbook

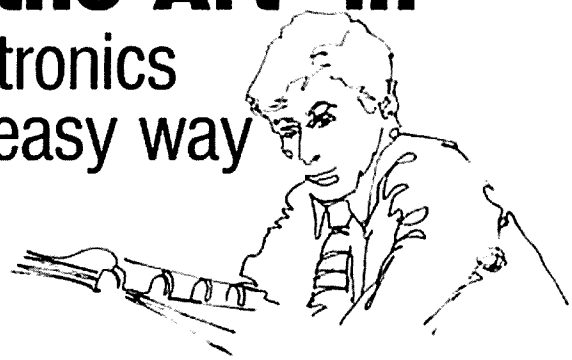
The complete story of the phase-locked loop is told in a free 76-page paperback entitled *Signetics Linear — Phase-Locked Loops Applications Book,* recently published by Signetics Corporation, a subsidiary of Corning Glass Works. The book is a companion to the larger *Linear Specifications Handbook* which is also available, although it must be requested separately, according to Jack Mattis, manager of consumer product marketing in the company's linear department.

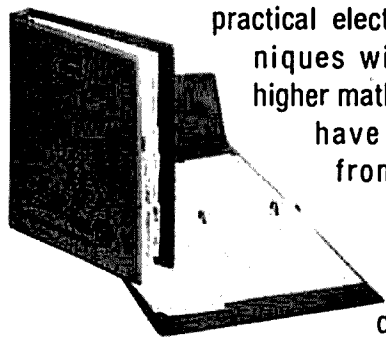
Phase-locked loops are a new class of monolithic integrated circuits developed by the Signetics research and development department in 1969 and marketed by the firm during the following year. They are based on frequency feedback technology which dates back 40 years. A phase-locked loop is basically an electronic servo loop consisting of a phase detector, a low-pass filter and a voltage-controlled oscillator. The controlled oscillator phase enables the PLL to lock or synchronize with an incoming signal.

In addition to the dash of history given in the book's introduction, other sections provide a short glossary and descriptions of the phase-locked loop principle and PLL "building blocks." Major sections include explanations of general loop setup and tradeoffs, PLL measurement techniques, monolithic phase-locked loops, expanding loop capability, and specific applications.

Some of the more interesting passages contain information on how the user of the PLL can apply the circuit to his own projects. As a functional building block, the phase-locked loop is suitable

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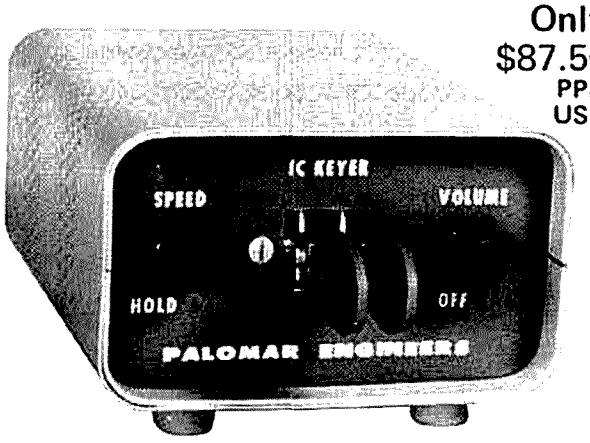
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for a wide variety of frequency-related applications. These generally fall into one or more of the following categories: fm demodulation, frequency synthesis, frequency synchronization, signal conditioning, and a-m demodulation. Each category is covered by a section in the book.

A number of construction projects are suggested as a means of proving the feasibility of using the phase-locked loop circuit in specific applications. The book provides information on building an fm i-f amplifier and demodulator, a phase-locked a-m receiver, an i-f stage with agc and a-m/fm detection, a translation loop for precise fm i-f generation for tv, a phase-locked FSK demodulator and many others. For a free copy of the handbook, write to Signetics PLL Handbook, Signetics Corporation, 811 East Arques Avenue, Sunnyvale, California 94086.

integrated-circuit fm detector

A unique method of fm detection by a new technique of linear gating is featured in the new Signetics ULN2111 monolithic integrated circuit. This linear device comprises a three-stage limiter and a balanced product detector.

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For more information, write to Signetics Corporation, 811 East Arques Avenue, Sunnyvale, California 94086.

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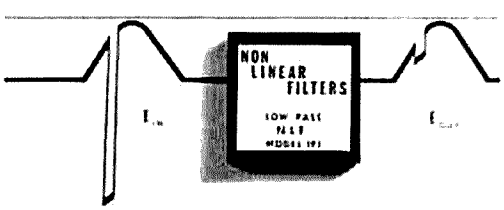
from cb to ham beginner

Whether you are a CBer or an SWL – or anyone else who wants to get started in amateur radio – this book answers your questions. It is an easy-to-read book that explains how to select and tune a communications receiver, listen in on the ham bands, acquire technical knowledge by building and experimenting, learn to send and receive code, build or buy an amateur transmitter, erect an effective antenna, and put an amateur station on the air.

In addition, the author, J.A. Stanley, presents a *fun* way of obtaining your Novice license. Instead of spending countless hours reading dry theory, simple tests are provided that use low-cost, readily available parts. A transmitter project is even included, so you can get on the air as soon as you get your license.

The book is filled with photos and drawings to simplify the subject matter. You need no technical background other than that obtained from operating a CB rig to understand and enjoy this book. It's an invaluable reference source that tells you how to become a Novice-class radio amateur. All the latest FCC Novice rule and frequency changes are included. 144 pages, softbound. $4.25 from Comtec Books, Greenville, New Hampshire 03048.

non-linear filters



Non Linear Filters has introduced a non-linear low-pass filter module that attenuates frequencies above the corner frequency while introducing no phase shift, either above or below the corner

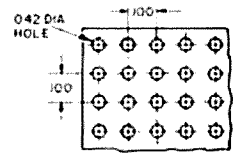
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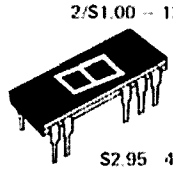


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frequency. This filter is significantly more effective than linear filters for eliminating noise spikes. For example, high amplitude noise spikes occurring on fm carriers are highly attenuated before limiting. Since the non-linear low-pass filter introduces no phase delay distortion, it is useful where non-sinusoidal waveshapes are encountered, such as in wideband digital communications systems (square waves), spectroscopy (Gaussian-shaped peaks) and narrow-band telephone-line digital communications.s,

The non-linear low-pass filter is effective in systems that have undesirable commutating spikes such as samplers, encoder and time-shared demultiplexers. The filter can also be used to filter wide-band white noise without introducing any phase shift in the signal of interest.

The corner frequency of the non-linear low-pass filter module is set by the user by adding two external capacitors to the unit. The corner frequency can be set anywhere in the range from dc to 10 kHz. For more information and applications data, write to Non Linear Filters, Box 338, Trumbull, Connecticut 06611, or use *check-off* on page 110.

semiconductor replacement manual

A comprehensive 52-page Semiconductor Replacement Manual has just been released by the Sprague Products Company. Containing over 30,000 OEM part numbers, listed alpha-numerically, which can be replaced by Sprague's new line of 82 popular semiconductor devices, this manual also includes performance characteristics, outline drawings and pertinent parameters for the entire Sprague line.

Included in the new Sprague semiconductor replacement line are 42 small-signal and power transistors, 5 field-effect transistors, 6 silicon rectifiers, 14 linear integrated circuits and 15 LED devices.

These semiconductor devices provide exact replacements for components found in home/mobile entertainment and communications equipment. In addition,

More Details? CHECK–OFF Page 110

the LED devices (which include visible light sources, infrared light sources, alpha-numeric displays and opto-electronic photo transistors) offer hobbyists new options and designs in electronic experimentation.

Copies of Semiconductor Replacement Manual, K-500, may be obtained without charge from Sprague distributors, by writing to Sprague Products Company, Marshall Street, North Adams, Massachusetts 01247, or by using *check-off* on page 110.

course in radio fundamentals

The new fifth edition of *A Course In Radio Fundamentals* is a completely-rewritten version of this long-time favorite, now in its 30th year. Modernizing the text, plus the introduction of much new material to increase the scope, has almost doubled the previous size of the book. If you are introducing someone new to amateur radio, this is a good book to start him off with.

Unlike the preceding printings, which used *The Radio Amateur's Handbook* as a text, the present volume is a complete and independent study manual. Paced at an intermediate technical level, the treatment is quantitative to the extent permitted by restricting mathematics to simple algebra.

The twenty-six chapters fall into two categories: electrical and radio circuit fundamentals and basic principles of electronic amplifying devices. Each chapter ends with a collection of questions and problems for testing the reader's understanding of the subject discussed. Answers to problems are given at the end of the book. The study text is followed by ten experiments using inexpensive, readily-available equipment. Detailed experimental data are contained in each.

The new soft-cover edition has 184 pages and contains well over 200 illustrations, including circuits and graphs. Priced at $2.00 from your local electronics supplier, or order from Comtec Books, Greenville, New Hampshire 03048.

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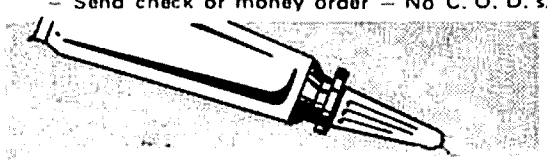
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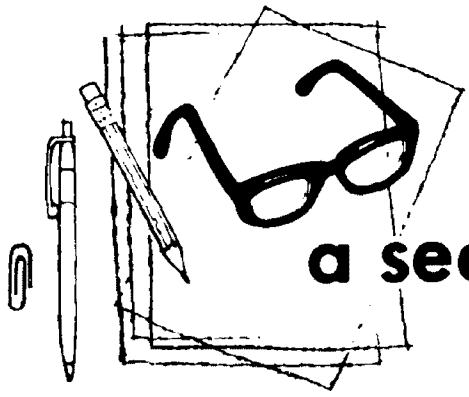
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a second look

by jim fisk

November 27th *marks* the fiftieth anniversary of one of amateur radio's most memorable events — the first two-way amateur communications across the Atlantic Ocean. It was a hard-won goal, its path marked with failure and frustration, but when the Atlantic, at last, had been spanned, it was conquered by *short-wave* amateur radio, on wavelengths that previously were considered to be useless.

The first Transatlantic tests, in December, 1920 were a dismal failure, as were a second series of tests conducted in February, 1921. The 250 or so British stations which were listening for prearranged signals from the United States on a wavelength of 200 meters jammed each other so badly with radiations from their own regenerative receivers that they couldn't hear any signals from across the pond!

A third Transatlantic test was scheduled for December, 1921. In November, Paul Godley, 2XE, designer of the famous Paragon receiver, sailed from New York with two receivers under his arm — one a standard variometer regenerative set with two stages of audio amplification, the other a 10-tube superheterodyne built especially for the tests. With this superhet and a Beverage antenna installed on the bleak Androssan moor on the coast of Scotland, Godley heard the first stateside signals coming through in the wee hours of the morning on December 8th.

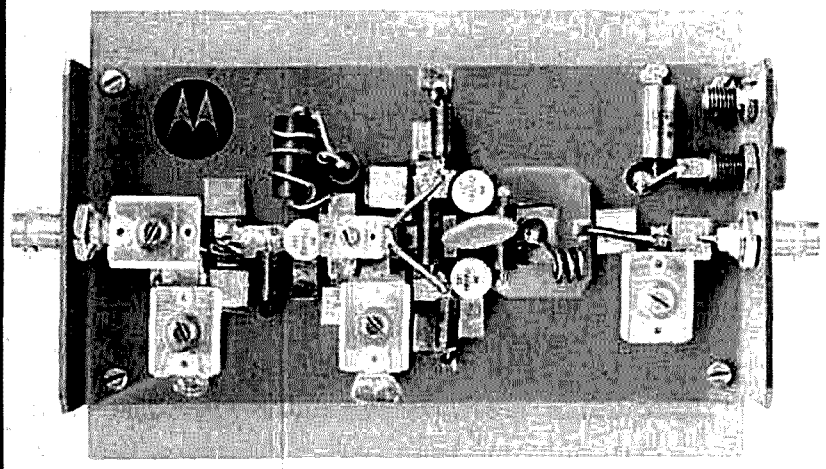
A year later, two European stations, F8AB in Nice, and G5WS in London, were heard along the east coast of the United States, but two-way communications were as elusive as ever.

A fourth series of Transatlantic tests were scheduled for late 1923. However, these carefully laid plans were totally upset by the enterprise of one man, Leon Deloy, F8AB. Deloy came to the states during the summer of 1923 where he met with John Reinartz, 1XAM, and Fred Schnell, 1MO. Deloy picked up a lot of valuable advice from his talks with Reinarts and Schnell, and before returning to France he acquired a new Grebe receiver and the details of a "trick" circuit which, he was told, would "go down to about 100 meters." Up until that time all the Transatlantic tests had been conducted on a wavelength of 200 meters.

Deloy put his new 100-meter station on the air in early autumn, and having satisfied himself that everything was in working order, cabled Schnell that he would transmit on 100 meters between 0200 and 0300 GMT on November 26, 1923. The signals from F8AB were heard by Schnell and Reinartz almost from the first dot he transmitted, but the Americans were not ready to transmit back. Unlike Deloy, who presumably did not think it was necessary to obtain official permission to transmit on such a short wavelength, Schnell had to seek the necessary authority from the Radio Supervisor in Boston.

On November 27th Schnell received special permits from Boston for himself and Reinartz. Late that night (early morning in Europe) they were both on the air. For an hour Deloy called the United States and then sent two messages. At 0330 GMT he signed off and asked for acknowledgement. Long calls followed from 1MO and 1XAM. Then came the eagerly awaited reply — Deloy had heard both stations clearly. Reinartz was asked to stand by as Deloy transmitted to Schnell, "R R QRK UR SIGS QSA VERY ONE FOOT FROM PHONES ON GREBE FB OM HEARTY CONGRATULATIONS THIS IS A FINE DAY — PSE QSL. It was, indeed, a fine day.

Jim Fisk, W1DTY
editor



a solid 80 watts for two meters

Construction details for high-performance, solid-state two-meter power amplifiers

John Hatchett, Motorola Semiconductor Products, Inc.

Evaluation results, component layout and construction information for two 80-watt vhf power amplifiers are described. These solid-state amplifiers can be used to boost two-meter output power levels to 80-watts. Both units have been designed to operate from a dc supply voltage of 12.5 volts with 50-ohm source and load impedances. The 12.5-volt power requirements are easily adapted to fixed or mobile operation.

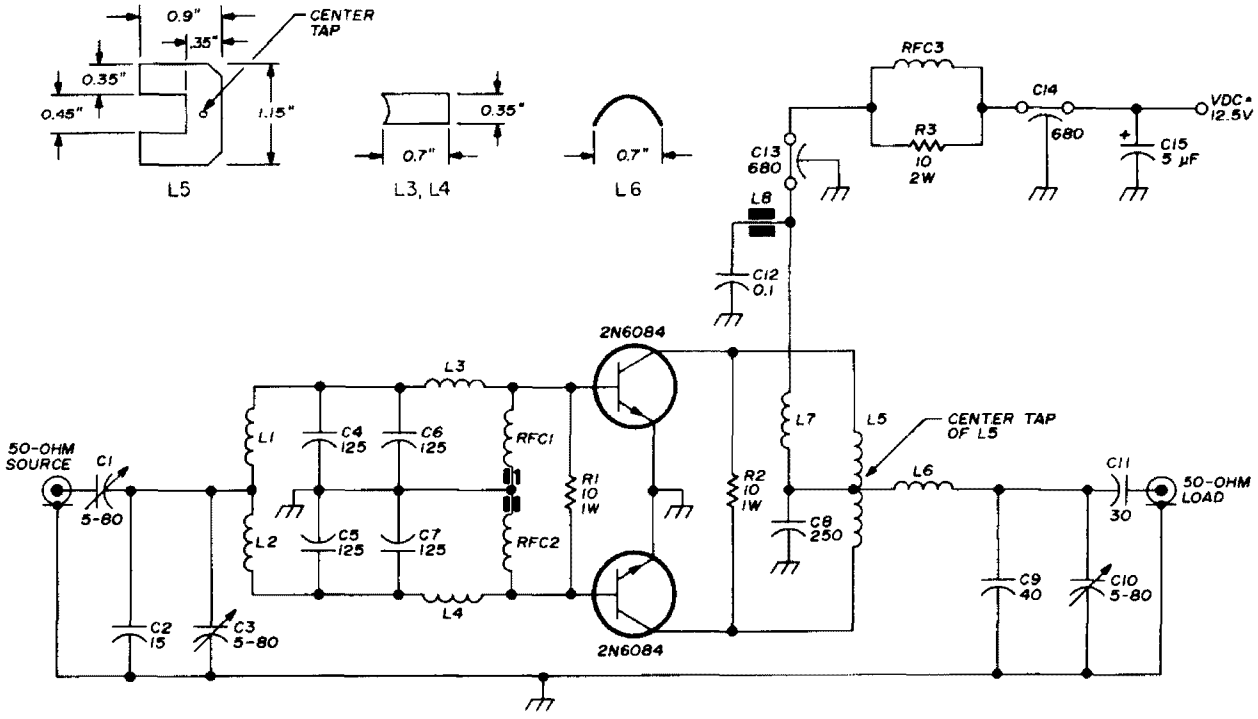
One of the amplifiers is a single-stage design using two 2N6084 transistors combined with simple LC components (fig. 1). It can be tuned to operate from 144 to 175 MHz, and requires a typical input power level of 20 watts for 80-watts output at 144 MHz.

The second amplifier uses the same output stage design, but adds a 2N6083 transistor driver stage (fig. 2) to reduce input drive requirements. This design is also tunable from 144 to 175 MHz, and will provide 80-watts output power at 144 MHz with only 2.5-watts of drive.

Six single- and two-stage amplifiers have been constructed and evaluated with similar performance exhibited by the amplifiers in each group. Typical values for the more important amplifier characteristics are shown in table 1 and in figs. 3, 4, 5 and 6. The amplifiers have also been subjected to momentary open- and short-circuit load conditions without damage to the transistors.

design philosophy

The amplifiers have been designed to be efficient, reliable and stable without sacrificing simplicity. All amplifier stages are of the common-emitter configuration, operated class-C. Two 40-watt 2N6084 transistors have been used in the high-power output stage to provide excellent heat distribution at full power. Combining the two 2N6084 devices is accomplished with LC signal splitting/combining techniques. For the single-stage amplifier the combinations of L1 and L3 and L2 and L4 split the signal, and inductance L5 recombines the signals. The two-stage amplifier uses L4 and L5 for signal splitting and L6 for combining. These inductors provide impe-



C1,C3,C10	5-80 pF trimmers (ARCO 462)
C2	15 pF metal clad (Underwood Electric type J-101*)
C4,C5,C6,C7	125 pF metal clad (Underwood Electric type J-101)
C8	250 pF metal clad (Underwood Electric type J-101)
C9	40 pF metal clad (Underwood Electric type J-101)
C11	30 pF metal clad (Underwood Electric type J-101)
C12	0.1 µF, 75 V ceramic disc
C13,C14	680 pF feedthrough (Allen Bradley type FA5C)
C15	5.0 µF, 25v, aluminum electrolytic
L1,L2	2½ turns number-16, 0.2" ID (60 nH)
L3,L4	cut from 0.031" single-sided G10 circuit board (5 nH)
L5	cut from 0.031" single-sided G10 circuit board (8 nH to center tap)
L6	number-12 wire, approximately 1.1" long (10 nH)
L7	3 turns number 14, 0.25" ID (50 nH)
L8	ferrite bead (Ferroxcube 5659065/3B)
RFC1,RFC2	0.15 µH molded choke with Ferroxcube 5659065/3B ferrite bead on ground lead
RFC3	10 turns number-14 wire wound around R3

fig. 1. Schematic diagram of the single-stage, 80-watt, 144-MHz power amplifier. Circuit is built on 0.062" single-sided G10 circuit board. Performance of this amplifier is graphed in figs. 3 and 5.

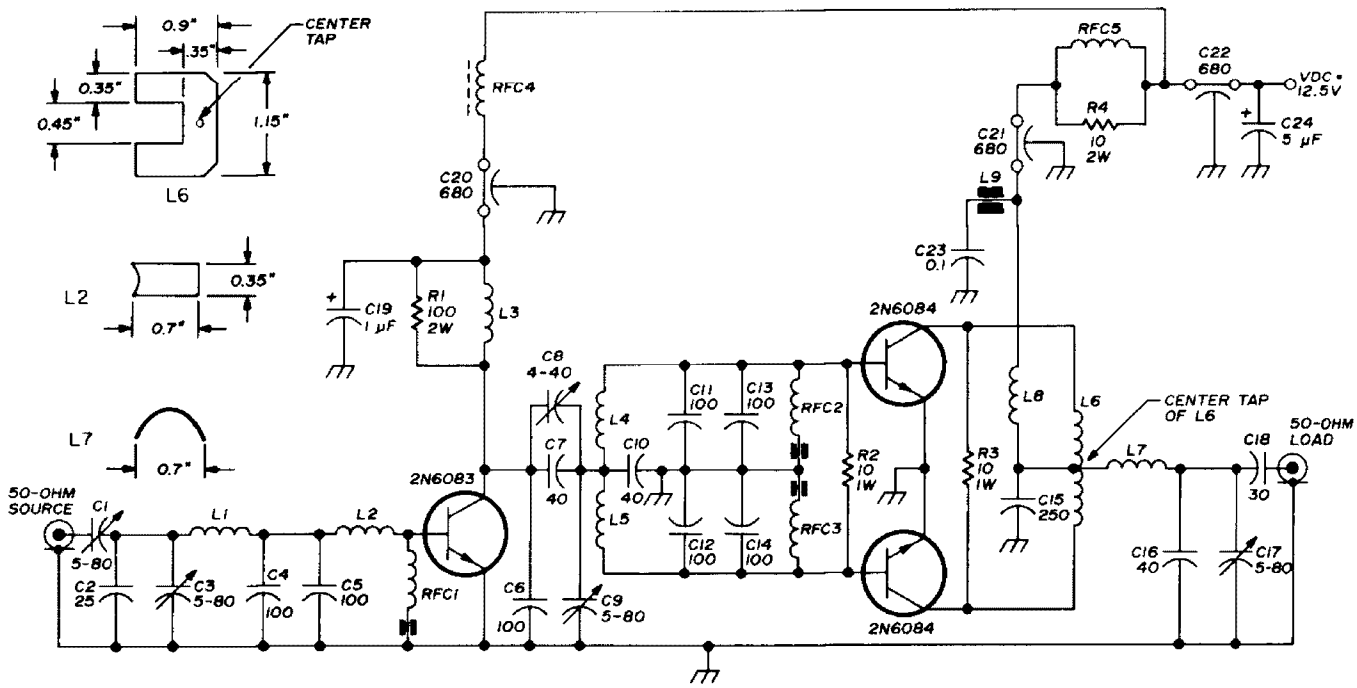
dance transformation, isolation between devices and minimize unequal load sharing.

Low-loaded Q impedance matching network designs have been used to maximize bandwidth and to minimize insertion loss. This also results in reducing reflected voltage levels that can occur during high-output vswr conditions. A low-pass, low-loss, LC output filter can be used to provide additional attenuation of the harmonic components.

The two transistor types used in the amplifiers are part of the Motorola vhf land mobile series designed for 12.5-volt fm operation. They are multiple balanced-emitter transistors manufactured using the †Isothermal process technology to minimize temperature variations across the transistor chips. This process provides increased transistor protection over wide thermal and load vswr excursions. The devices are packaged in a 0.380 inch diameter, stripline-opposed-emitter stud package (case 145A-01).

*Underwood Electric & Mfg. Co., Inc., 148 South Eighth Avenue, Maywood, Illinois 60153.

†Trademark of Motorola Inc.



C1,C3, C9,C17	5-80 pF trimmers (ARCO 462)
C2	25 pF metal clad (Underwood Electric type J-101)
C4,C5,C6, C12,C13,C14	100 pF metal clad (Underwood Electric type J-101)
C7,C10,C16	40 pF metal clad (Underwood Electric type J-101)
C8	4-40 pF trimmer (Arco 403)
C15	250 pF metal clad (Underwood Electric type J-101)
C18	30 pF metal clad (Underwood Electric type J-101)
C19	1.0 μF tantalum
C20,C21,C22	680-pF feedthrough (Allen Bradley type FA5C)
C23	0.1 μF, 75 V ceramic disc
C24	5.0 μF, 25V, aluminum electrolytic
L1	1 turn number-16, 0.25" ID (18 nH)
L3	3 turns number-16 wound around R1 (60 nH)
L4,L5	1.1" long number-14 wire, formed around 0.6" diameter cylinder (12 nH)
L2	cut from 0.031" single-sided G10 circuit board (5 nH)
L6	cut from 0.031" single-sided G10 circuit board (8 nH to center tap)
L7	number-12 wire, approximately 1.1" long (10 nH)
L8	3 turns number-14, 0.25" ID (50 nH)
L9	ferrite bead (Ferroxcube 5659065/3B)
RFC1, RFC2, RFC3	0.15 μH molded choke with Ferroxcube 5659065/3B ferrite bead on ground lead
RFC4	ferrite choke (Ferroxcube VK200 19/4B)
RFC5	10 turns number-14 wound around R4

fig. 2. Schematic diagram of the two-stage, 80-watt, 144-MHz power amplifier. Circuit is built on 0.062" single-sided G10 circuit board as shown in the photograph. Performance of this amplifier is shown in figs. 4 and 6.

To achieve the 80-watt power level, it is imperative that low-loss matching network components be used. It is also necessary that these components be characterized for the desired operating frequencies. Suitable low-loss coils can be made with a small length of wire, ribbon conductor or printed circuit board material. Economical capacitors for efficient high-power operation at 2 meters are more difficult to obtain. All fixed capacitors in the amplifiers, 250 pF or less in value, are Underwood mica dielectric units. The effective capacitance of these components at 2-meters will deviate only slightly from the low frequency value for nominal capacitance values up to approximately 60 pF. Larger capacitors of this type are characterized for operation at the selected frequency.

construction

A full description of all necessary components for building the amplifiers along with the schematic diagrams are shown in figs. 1 and 2. Care must also be given to the physical location of the components. The photograph and the scale drawings in figs. 7 and 8 can be used to determine proper component placement. For the sake of simplicity, only those components necessary to establish the basic amplifier layout have been included in the drawings.

table 1. Amplifier performance for a dc supply voltage of 12.5 volts.

		single-stage design	two-stage design
Power output (watts)		80	80
Power input (watts)	144 MHz	20	2.5
	148 MHz	21	2.6
	165 MHz	23	3.5
	175 MHz	26	5.5
Power gain at 144 MHz	(dB)	6.0	15.1
Dc current (amperes)	output stage	8.5	8.5
	driver stage	—	2.5
Harmonic attenuation	(dB)	20	20
Stability	Amplifiers are stable for input drive levels from zero to 30% overdrive and for supply voltages from 8.0 to 15.5 volts dc.		
Ruggedness	With 80 watts power output into 50 ohms, no transistor damage from open- and short-circuit load conditions for all phase angles		

The amplifiers are built on 0.062 inch, single-sided, G10 circuit board with the components mounted on the ground plane side. In each case, the ground plane is continuous except for interruptions for the transistor and feedthrough capacitor (C13, C14 and C20, C21, C22) mounting holes. Coils L3, L4 and L5 of the single-stage amplifier isolate the transistor base and collector contacts from the ground plane. Coils L2 and L6 accomplish this function in the two-stage design. In addition, four small pads of 0.31-inch, G10 circuit board are used to provide isolation for the 2N6083 collector, the base of each 2N6084 and capacitor C7.

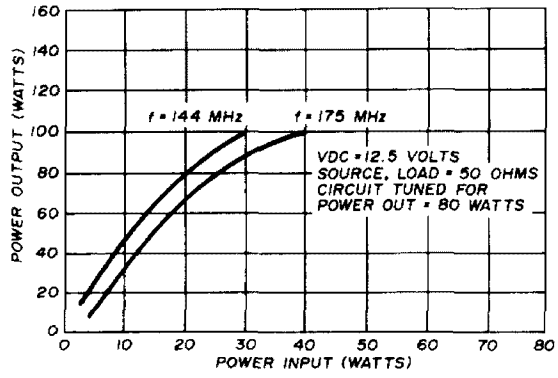


fig. 3. Power output vs power input for the single-stage amplifier.

To prevent physical damage to the transistor stud package, the following precautions should be observed:

A. The maximum torque ratings for the mounting nut must not be exceeded (6.5 inch-pounds for the 2N6083 and 2N6084 devices).

B. The nut should be placed on the stud and tightened to the specified torque before soldering the transistor leads to the circuit. After the nut is properly torqued, a slightly downward pressure can be exerted on the leads to place them in contact with the circuit board connection points. The objective is to prevent an upward force being applied to the leads near the case body.

thermal considerations

The amplifiers must be provided with

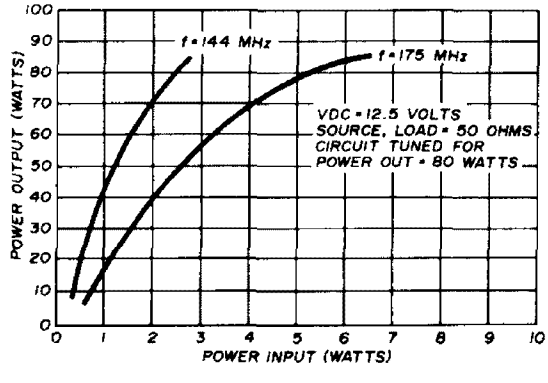


fig. 4. Power output vs power input for the two-stage amplifier.

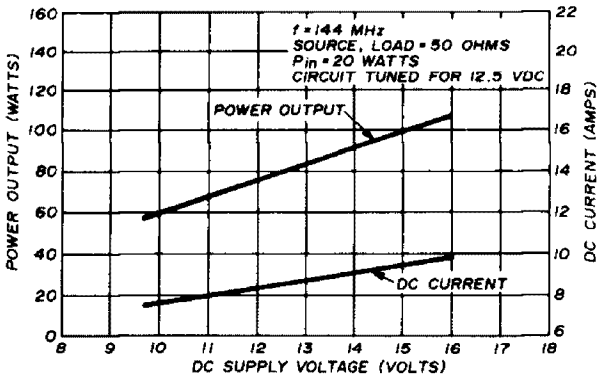


fig. 5. Amplifier power output and dc current vs dc supply voltage for the single-stage amplifier.

heat sinks capable of keeping the transistor junction temperatures below their specified maximum temperature of 200°C. This requires extremely good thermal design and construction practice. A smooth heat-sink surface is required to maximize heat-sink to transistor case contact area. A proper amount of thermal-joint compound must be used between heatsink and transistor case interface to improve thermal transfer to the heatsink. Wakefield type 120, Thermoalloy *Thermacote,* Dow Corning type 340 or other thermal compounds exhibiting similar low thermal resistance properties are recommended. The heatsink must have a thermal resistance low enough to adequately transfer the heat from the transistor case to surrounding air.

Limiting the transistor junction temperatures to a maximum of 180°C during continuous operation into a 50-ohm load requires a heat-sink thermal resistance specification of less than 1.7°C/watt for the output stage devices at 60°C ambient. For an ambient of 30°C, the heat-sink thermal resistance requirement can be relaxed to approximately 2.3°C/watt. Similar operating conditions require the 2N6083 driver transistor heat-sink thermal resistance to be less than approximately 6° and 8°C/watt for ambient temperatures of 60° and 30°C, respectively.

Duty cycle operation, such as 1-minute on/3-minutes off, will significantly reduce the heat-sinking requirements. If operation into mismatched loads is anticipated, the heat-sink thermal resistance values must be reduced to account for the radical increase in transistor power dissipation that can occur with these operating conditions.

Several economical aluminum heatsinks are available with thermal resistance values in the order of 3°C/watt. These would be adequate for use with the amplifiers in most applications, since a 50-ohm load is used and continuous operation capability is not required. More expensive heatsinks can provide thermal resistance values less than 1°C/watt. **Table 2** provides a brief description for some of the commercially available units.

amplifier adjustment

An amplifier alignment test set-up is shown in **fig. 9.** Initial amplifier tuning should be started with reduced supply voltage (approximately 8 volts) and reduced drive levels to prevent excessive device dissipation. For 144-MHz operation, a reasonably good starting point would be to set all variable capacitors approximately ½-turn from the fully closed (maximum capacity) position. During alignment, you may carefully touch each transistor case to detect excessive power dissipation in any of the transistors. Each transistor case should feel warm, but not too hot.

If a spectrum analyzer is available, it should be used to monitor the output signal during tuneup to verify proper alignment and to indicate the presence of low-frequency oscillations that can occur if the amplifiers are significantly mal-

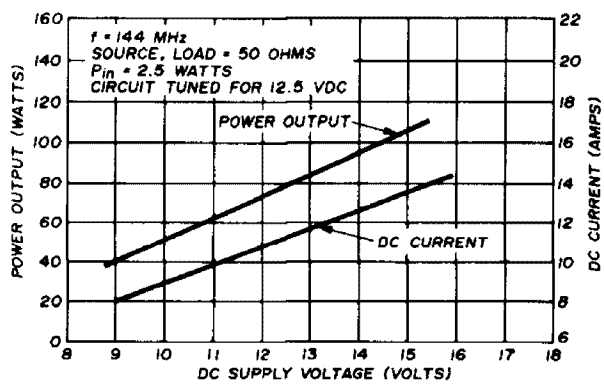


fig. 6. Amplifier power output and dc current vs dc supply voltage for the two-stage amplifier.

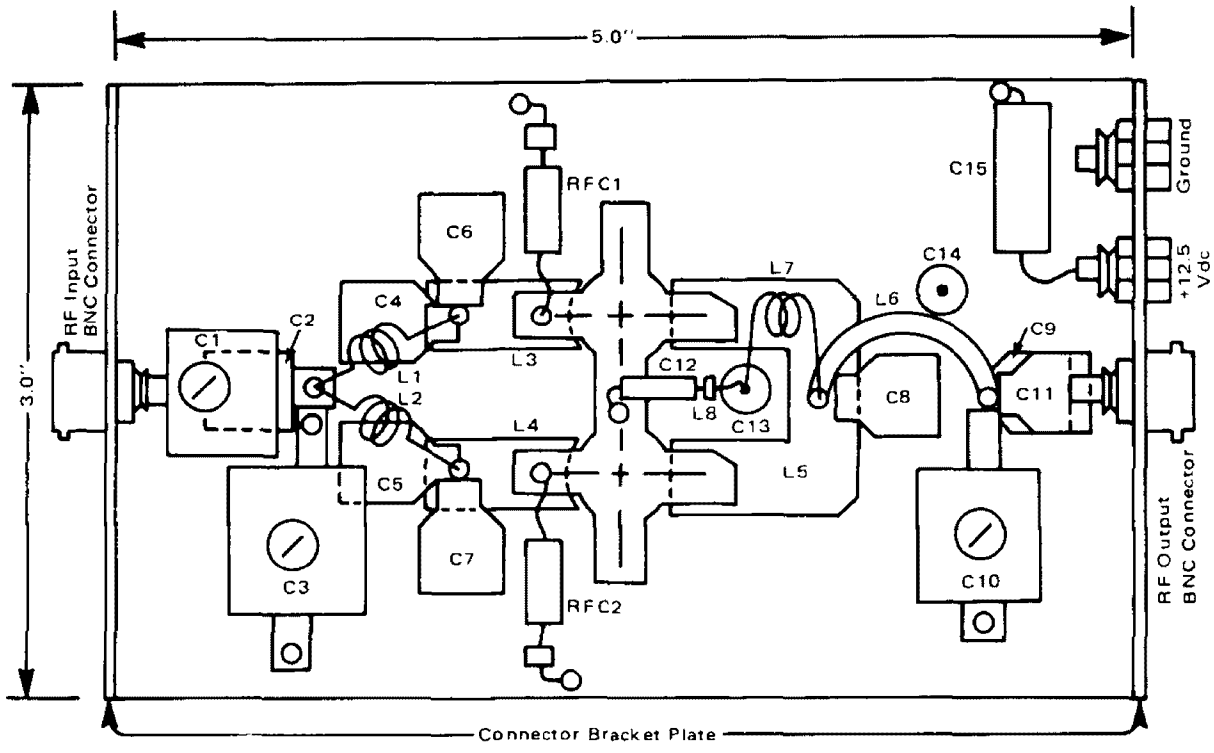


fig. 7. Component location for the single-stage amplifier. Some components have been omitted for clarity. Transistor mounting holes are 0.80", center to center. RFC3 and R3 are mounted on rear side of board.

adjusted. An oscilloscope connected to the dc voltage line (for example, at the top of the 5-μF filter capacitor, C15 or C24) can also be used to provide useful information on the presence of low-frequency oscillations. The scope probe will usually pick up enough of the two-meter signal energy to provide a signal display on the CRT.

If low-frequency oscillations are not present, the two-meter signal display will be constant in amplitude. If a low-frequency oscillation (typically less than 10 MHz) is present, it will show up as amplitude variations on the two-meter display. The frequency of the amplitude variations correspond to the frequency of the oscillation. High-frequency oscilloscopes (100 MHz) will provide a good display of the two-meter signal. Low-frequency oscilloscopes (20 MHz) are not capable of showing the two-meter signal itself, but they can be useful in determining if a low-frequency amplitude variation (envelope) is present on the two-meter signal. Any oscillation should be eliminated by adjusting the amplifier variable capacitors.

Single-stage amplifier. Start with low drive level (approximately 2 to 5 watts)

table 2. Summary of commercial heatsinks suitable for use with the 80-watt, two-meter amplifier. Thermal resistance values (column 2) are for natural convection except for the Wakefield FC-502 and FC-503 units, which are for 10 cubic feet/minute air flow.

part number	approximate thermal resistance (°C/watt)	description
WEI Corp. 3110	2.5	aluminum, 1.3" x 4.0" x 1.5" & 3.0"
WEI Corp. 3164	2.5	aluminum, 1.0" x 4.12" x 1.5" & 3.0"
Thermalloy 6169	2.5	aluminum, 1.3" x 4.12" x 3.0"
Wakefield NC-641	2.5	aluminum, 1.0" x 4.12" x 3.0"
Wakefield FC-502	0.45	copper, 1.75" x 3.5" x 1.75"
Wakefield FC-503	0.35	copper, 1.75" x 3.5" x 3.5"

WEI Corporation, P. O. Box 10577, Santa Ana, California 92705
Thermalloy Inc., 8717 Diplomacy Row, Dallas, Texas 75247
Wakefield Engineering Inc., Wakefield, Massachusetts 01881

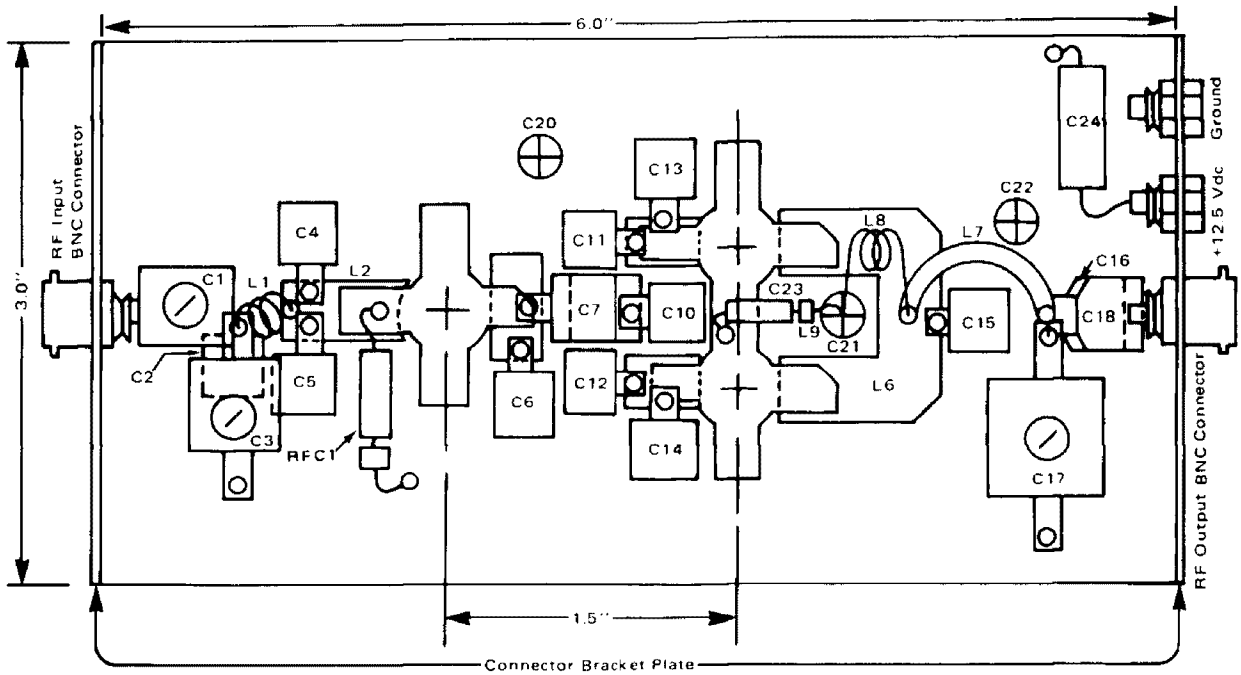


fig. 8. Component location for the two-stage amplifier. Some components have been omitted for clarity. Transistor mounting holes are 0.80", center to center. RFC4, RFC5 and R4 are mounted on back of board.

large enough to turn on the transistors (indicated by the flow of dc collector current). Then adjust capacitor C10 for maximum output and C1 and C3 for minimum reflected power to the drive source as indicated by the swr bridge. Increase the supply voltage to 12.5 volts after this initial adjustment, and continue to increase the input drive power to approximately 8 to 12 watts while adjusting C10 first and then C1 and C3 as before.

Increase the drive power to approximately 20 watts, and tune for rated power output in a similar manner. After tuning for rated output power, capacitor C10 can be increased slightly in capacitance. This will minimize the required dc current with only a slight degradation, approximately 0.1 dB or less, in power output.

Two-stage amplifier. Start with low drive (approximately 0.25 to 0.5 watt) large enough to turn on the 2N6083 stage as indicated by the flow of dc collector current. Then adjust C8, C9 and C17 for maximum output power and C1 and C3 for minimum reflected power to the drive source as indicated by the swr bridge. Increase the supply voltage to 12.5 volts after this initial adjustment, and increase the input drive power to approximately 1.0 to 1.5 watts while adjusting C17 first and then C8, C9 and C1, C3 as before.

Now, increase the drive power to approximately 2.5 watts and tune for rated power output. After tuning for rated output power, capacitor C17 can be increased slightly in capacitance. This will minimize the output stage dc current requirement with only a slight degradation, approximately 0.1 dB or less, in power output.

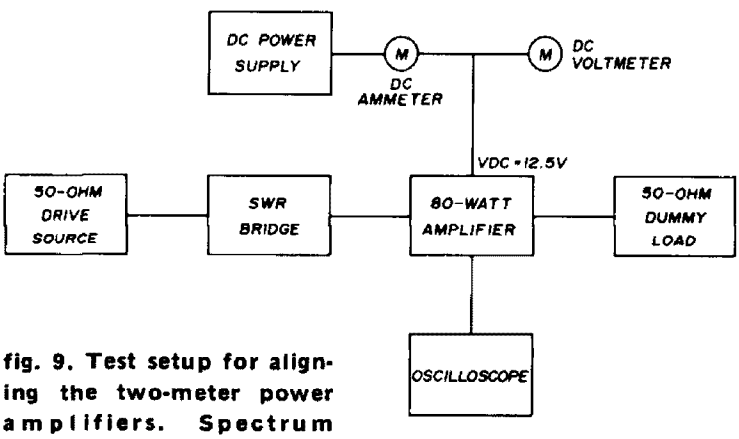
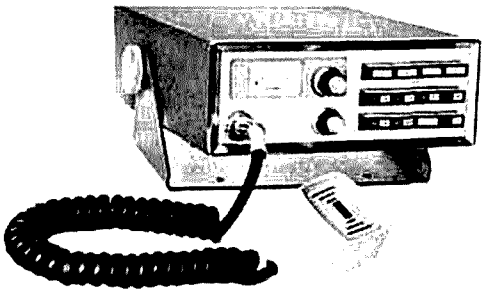


fig. 9. Test setup for aligning the two-meter power amplifiers. Spectrum analyzer and power attenuators should be used if available.

ham radio



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It's an all solid-state design that you can build and completely align without special instruments. And this compact little beauty gives you 36 channel capability with independent push-button selection of 6 transmit and 6 receive crystals. 10 watts minimum output into an infinite VSWR without failure. And for the ultimate in convenience there's the optional tone burst encoder for front panel selection of four presettable tones. The HW-202 kit includes two crystals for set-up and alignment and simplex operation on 146.94; push-to-talk mike; 12-volt hook-up cable; heavy duty clips for use with temporary battery; antenna coax jack; gimbal bracket, and mobile mounting plate.

HW-202 SPECIFICATIONS—RECEIVER—Sensitivity: 2 dB SINAD* (or 15 dB of quieting) at .5μv or less. **Squelch threshold:** 3μv or less. **Audio output:** 2 W at less than 10% total harmonic distortion (THD). **Operating frequency stability:** Better than ±.0015%. **Image rejection:** Greater than 55 dB. **Spurious rejection:** Greater than 60 dB. **IF rejection:** Greater than 75 dB. **First IF frequency:** 10.7 MHz ±2 kHz. **Second IF frequency:** 455 kHz (adjustable). **Receiver bandwidth:** 22 kHz nominal. **De-emphasis:** —6 dB per octave from 300 to 3000 Hz nominal. **Modulation acceptance:** 7.5 kHz minimum. **TRANSMITTER — Power output:** 10 watts minimum. **Spurious output:** Below —45 dB from carrier. **Stability:** Better than ±.0015%. **Oscillator frequency:** 6 MHz, approximately. **Multiplier factor:** X 24. **Modulation:** Phase, adjustable 0-7.5 kHz, with instantaneous limiting. **Duty cycle:** 100% with ∞ VSWR. **High VSWR shutdown:** None. **GENERAL — Speaker impedance:** 4 ohms. **Operating frequency range:** 143.9 to 148.3 MHz. **Current consumption: Receiver (squelched):** Less than 200 mA. **Transmitter:** Less than 2.2 amperes. **Operating temperture range:** —10° to 122° F (—30° to + 50° C). **Operating voltage range:** 12.6 to 16.0 VDC (13.8 VDC nominal). **Dimensions:** 2¾" H x 8¼" W x 9⅞" D.

*SINAD = $\frac{\text{Signal + noise + distortion}}{\text{Noise + distortion}}$

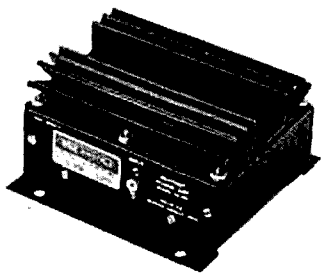
Kit HW-202, 11 lbs., mailable	179.95*
Kit HWA-202-2, Tone Burst Encoder, 1 lb.	24.95*
Kit HWA-202-1, AC Power Supply, 7 lbs.	29.95*
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. . . and here's 40 watts out for your 10 watts in

The Heathkit HA-202 2-Meter Amplifier works with any 2-meter exciter delivering 5-15 watts while pulling a meager 7 amps from any 12 VDC system. No additional power supplies are required. All solid-state components mount on a single circuit board for easy two-evening assembly. Manual shows exact alignment procedures using a VOM or VTVM. Connecting cable and antenna cable are included.

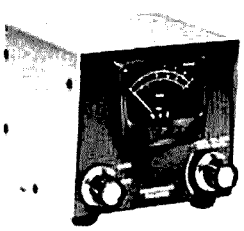
Kit HA-202, 4 lbs. 69.95*

HA-202 SPECIFICATIONS — Frequency range: 143-149 MHz. **Power output:** 20W @ 5 W in, 30W @ 7.5W in, 40W @ 10 W in, 50W @ 15 W in. **Power input (rf drive):** 5 to 15W. **Input/output impedance:** 50 ohms, nominal. **Input VSWR:** 1.5:1 max. **Load VSWR:** 3:1 max. **Power supply requirements:** 12 to 16 VDC, 7 amps max. **Operating temperature range:** —30° F. to +140° F. **Dimensions:** 3" H x 4¼" W x 5½" D.



$69.95*

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The Heathkit VHF/SWR Bridge tests transmitter output in power ranges of 1 to 25 watts and 10 to 250 watts ± 10% of full scale. 50 ohm nominal impedance permits placement in transmission line permanently with little or no loss. Built-in SWR bridge for tuning 2-meter antenna for proper match, has less than 10-watt sensitivity.

Kit HM-2102, 4 lbs. 29.95*

HM-2102 SPECIFICATIONS — Frequency range: 50 MHz to 160 MHz. **Wattmeter accuracy:** ±10% of full-scale reading.* **Power capability:** To 250 W. **SWR sensitivity:** less than 10 W. **Impedance:** 50 ohms nominal. **SWR bridge:** Continuous to 250 W. **Connectors:** UHF type SO-239. **Dimensions:** 5¼" W, 5¼" H and 6½" D, assembled as one unit.
*Using a 50 Ω noninductive load.

$29.95*

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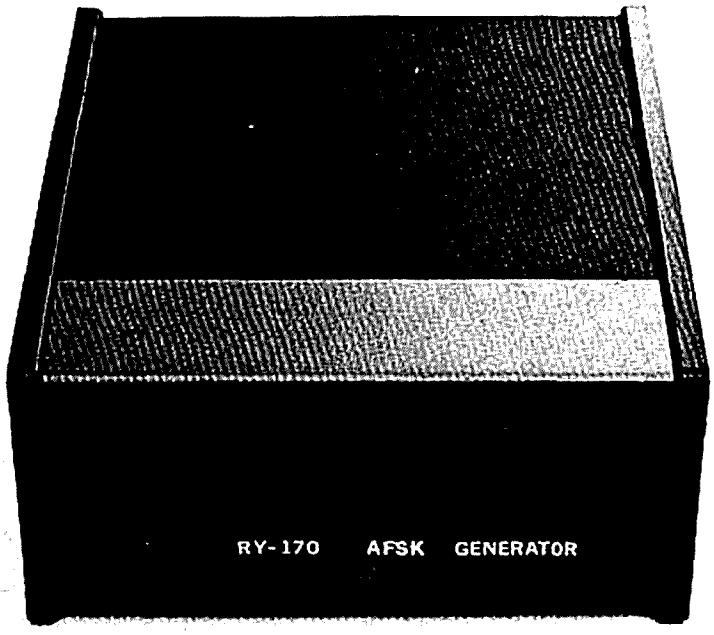
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crystal controlled AFSK generator

Complete construction details for the RY-170 — an AFSK synthesizer for 170-Hz shift

Howard Nurse, W6LLO, Palo Alto, California 94306

How would you like to generate precise RTTY audio tones without the need for a counter to establish the correct frequencies? Many years ago, when faced with the same problem, I recall an attempt to use a guitar to help adjust an AFSK oscillator! With that technique leaving something to be desired, I often thought how nice it would be to have an oscillator which could generate the correct frequencies without adjustment, while not breaking the bank in the process. Enter the RY-170, described here.

It wasn't until recently that surplus integrated circuits have made inexpensive frequency synthesis techniques possible. Start with a surplus crystal, divide by the correct ratios to generate 2125 and 2295 Hz, add a simple active bandpass filter, and for less than ten dollars you can have a 170-Hz shift synthesizer in your RTTY system.

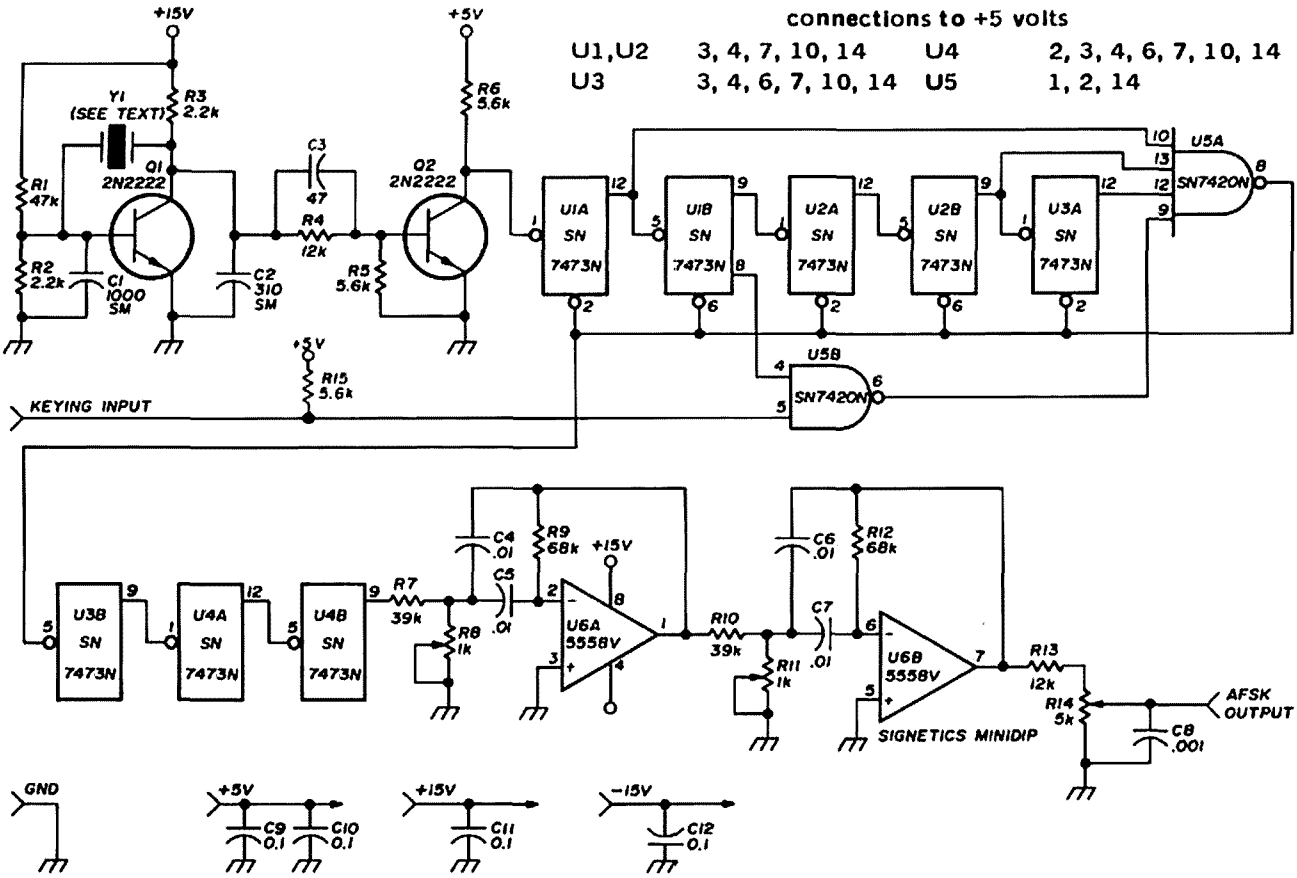


fig. 1. Schematic diagram of the RY-170 AFSK generator. Pin 11 of ICs U1, U2, U3 and U4, and pin 7 of U5, are connected to ground.

design

The following goals were established prior to starting the RY-170 project: Generation of 2125- and 2295-Hz tones from one crystal, low output distortion (THD), minimal keying overshoot, control from a TTL compatible input, use of inexpensive components, and easily duplicated printed circuit board. The resulting design is shown in fig. 1, while the photographs show various views of the completed unit. A summary of the RY-170 specifications is given in table 1.

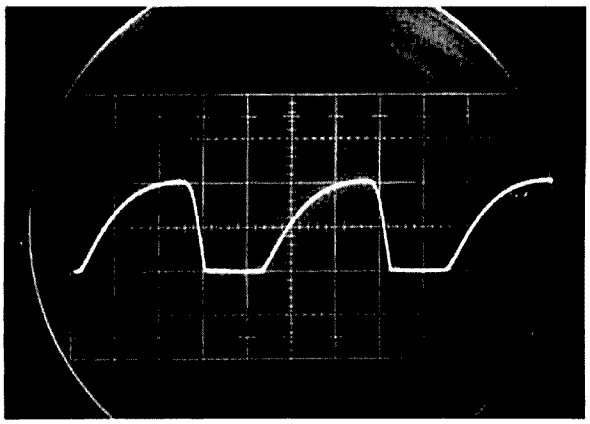
circuit description

The heart of the AFSK synthesizer is an oscillator using a surplus FT-241 crystal and transistor Q1 in a modified Pierce circuit. A channel 48 (459.259 kHz) crystal will yield output frequencies accurate to approximately 2 Hz, while preserving the relative shift (170 Hz) to within 0.1 Hz. If you desire even greater accuracy (with a slight increase in cost), an FT-241 crystal can be ordered which has been adjusted to the correct fre-

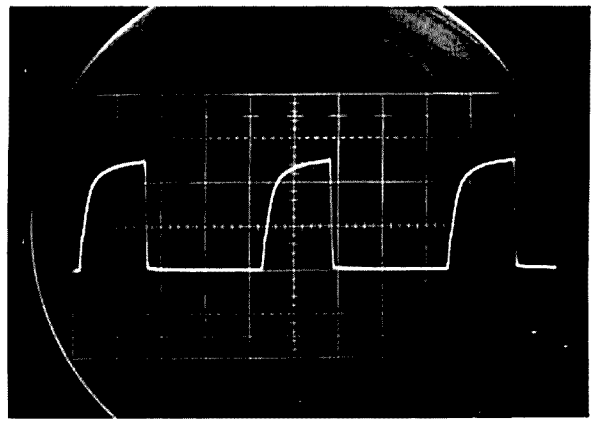


Rear view of the RY-170 AFSK generator with the top cover removed.

fig. 2. Waveforms of the RY-170 AFSK generator.



A. Collector Q1
horizontal scale, 0.5 microseconds/cm vertical scale, 5.0 volts/cm ground 2 cm from bottom

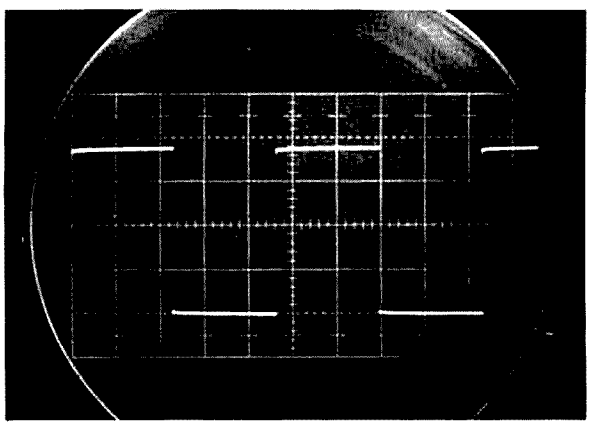


B. Collector Q2
horizontal scale, 0.5 microseconds/cm vertical scale, 2.0 volts/cm ground 2 cm from bottom

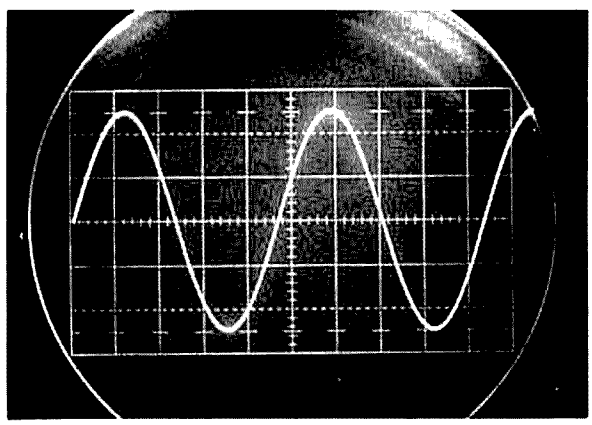
quency of 459.000 kHz.* The output waveforms from the oscillator and its buffer are shown in figs. 2A and 2B, respectively.

output waveform of the complete divider is a TTL square wave, shown in fig. 2C.

The output bandpass filter, necessary to extract the fundamental frequency

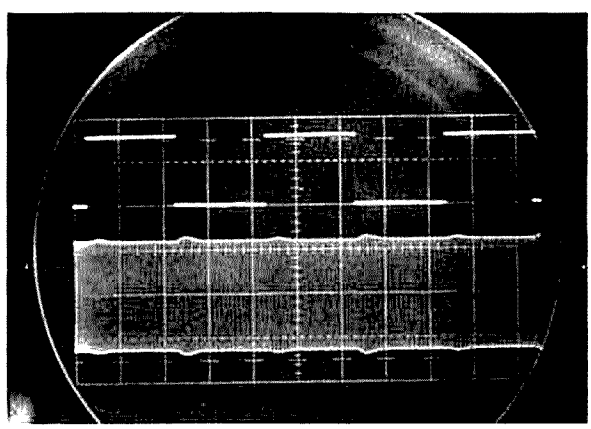


C. Divider output
horizontal scale, 100 microseconds/cm vertical scale, 1.0 volt/cm ground 1 cm from bottom



D. Filter output
horizontal scale, 100 microseconds/cm vertical scale, 0.5 volts/cm ground at center

Transistor Q2 interfaces the output of the oscillator to the divider input. JK flip-flops U1 through U3A and NAND gate U5 are wired as a programmable divider whose divide ratio depends on the logic state of the synthesizer input. When the input is grounded, the divide ratio is 25 (2295 Hz); when the input is high, the ratio is 27 (2125 Hz). The programmable portion of the divider is followed by an additional divide-by-eight circuit consisting of flip-flops U3B and U4. The



E. Top trace — TTL-compatible keying waveform horizontal scale, 5 milliseconds/cm vertical scale, 2 volts/cm
Bottom trace — RY-170 output horizontal scale, 5 milliseconds/cm vertical scale, 0.2 volts/cm

*JAN Crystals, 2400 Crystal Drive, Ft. Myers, Florida 33901.

from the divider output, consists of dual-operational amplifier U6 and its associated components. The filter was designed with a Q of 10; raising the Q would result in increased filter sensitivity to component tolerances and increased keying overshoot, while lowering the Q would raise the THD.

adjustments

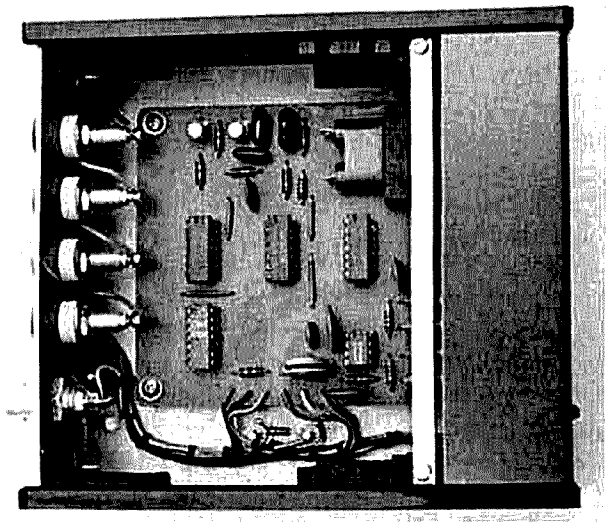
Resistors R8 and R11 are used to adjust the filter center frequency to pass the two tones. Resistors R8 and R11 should be adjusted, one for each tone, so that the transmitter has equal output power for 2125- and 2295-Hz inputs.The sinusoidal output waveform of the band-pass filter is shown in fig. 2D while the keying characteristics are shown in fig. 2E.

The output amplitude of the synthesizer can be adjusted with resistor R14 to match the transmitter audio requirements. R14 should be adjusted in conjunction with R8 and R11 to ensure that the audio stages of the transmitter are not overloaded.

Because the RY-170 is part of a larger system, I decided to use a common power supply for all accessories. If it is desired to use an internal supply with the AFSK board, a regulated supply meeting the requirements of the circuit can be used (see specifications in table 1).

table 1. Specifications for the RY-170 170-Hz AFSK synthesizer.

Absolute frequency	459.000-kHz crystal: Mark: 2125 ±1 Hz Space: 2295 ±1 Hz Channel-48 crystal: Mark: 2126 ±1 Hz Space: 2296 ±1 Hz
Frequency shift	170.0 ±0.1 Hz
Output amplitude	0-1 volt p-p
Output distortion (THD)	less than 0.5%
Output keying overshoot	less than 5%
Power requirements	+5 ±0.5 volts at 100 mA +15 ±1 volts at 10 mA -15 ±1 volts at 10 mA
Keying	TTL high (open) for mark TTL low (ground) for space
Crystal frequency	459.000 kHz or Channel 48 FT-241
Divide ratio	Mark: 27 x 8 = 216 Space: 25 x 8 = 200



Component layout of the RY-170 AFSK generator. A printed-circuit board is available from the author.

construction

The RY-170 AFSK synthesizer was constructed in a Ten-Tec JG-5 enclosure. The front panel was painted with Krylon 2021 (Oldsmobile green) which closely matches the color of the Heath SB-series. An LED is used as a pilot light, powered from the +5-volt power supply through a 220-ohm current-limiting resistor. The front and back panels are labelled with press-on letters and sprayed with Datak *Datakoat* for protection.

The circuit board is single-sided G-10 board and requires four jumpers.* I used Molex connector pins to hold the ICs although the ICs can also be soldered directly to the printed-circuit board. Dipped mica capacitors are required in the oscillator circuit, while high-stability capacitors (Orange Drop or polystyrene) should be used in the active filter.

There is something satisfying in knowing your shift is, and will remain, at 170 Hz. The RY-170 is one answer, and an economical one at that, to stable 170-Hz AFSK shift.

ham radio

*Drilled circuit boards and component layout information are available from the author for $5.50, postpaid.

More Details? CHECK–OFF Page 126

wide range rf signal generator

This wide range signal generator covers the range from 600 kHz to 12 MHz and features a built-in 1-kHz modulator

Hank Olson, W6GXN, P.O. Box 339, Menlo Park, California 94025

Most signal generators used by amateurs and other radio experimenters use the LC tuned oscillator in one form or another. Whether the exact circuit is a Colpitts or Hartley oscillator or some variation of these two basic designs, the output frequency is usually proportional to the inverse of the square root of the capacitance of the main tuning capacitor. That is, at least approximately

$$f = \frac{1}{2\pi\sqrt{LC}}$$

Because of this relationship signal generator tuning is broken into bands, each of which encompasses a high- to low-frequency ratio of only 3 or 4 to one (check the dial on your own signal generator, and see). The 3 or 4 to one frequency ratio is a direct consequence of the fact that parallel-resonant LC circuits vary in frequency as $1/\sqrt{C}$, and most variable tuning capacitors have a maximum to minimum capacitance ratio of 20 to 1 or less.

The oscillator described here is *not* an LC type, and its output frequency is proportional to 1/C. Thus, the variable capacitor is capable of tuning the oscillator over a 20 to 1 range, from 600 kHz to 12 MHz! Since such a wide range of frequency is covered by a single 180° turn of the capacitor shaft, it is advisable to have a good sized dial for calibration. The largest dial of good quality that I could readily obtain (the biggest one in the junkbox) was a Millen 10035. The use of this dial is the only reason that the signal generator is as large as it is.

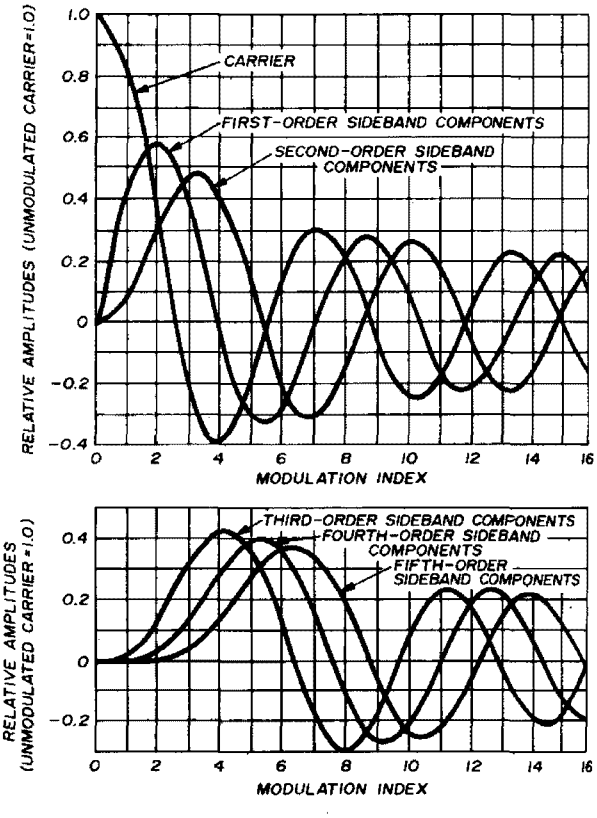


fig. 1. Graph of fm sideband components shows that the first-order sideband is nearly linear up to a modulation index of 0.5, and that higher-order sidebands are nearly non-existent at lower modulation indices.

circuit

A standard 365-pF broadcast tuning capacitor is used in the signal generator in conjunction with half of a relatively new Motorola IC, the Motorola MC4024P (or HEP 3805P). This IC is characterized as a dual voltage-controlled multivibrator, or vco. Since only half of the MC4024P is used to produce the rf output, the other half can be used to generate a 1-kHz modulation frequency.

The *waveforms* produced by both halves of the MC4024P are rectangular, and contain many harmonics. The harmonics of the rf oscillator section are not particularly troublesome, since *most* signal generators have appreciable harmonic content. However, it is desirable to filter the 1-kHz modulation waveform so that only one set of sidebands will be produced when the rf is modulated by the 1-kHz signal. Actually, since small index *frequency* modulation is used, there will be some small higher-order sidebands at 2 kHz and higher spacing around the rf carrier, but these will be insignificant if the generator is used within the bounds of narrowband frequency modulation (nbfm).

Nbfm has long since passed from the amateur scene, at least as a modulation method on the high-frequency bands. The main reason for nbfm's disfavor is that it is useful only for simulating amplitude modulation with 50% or smaller percentages. If higher indexes of fm are used, the higher-order sidebands rapidly increase and the signal no longer resembles a-m.

To see how this works, look at the graphs of **fig. 1**. Note that the graph representing the first-order sidebands is approximately *linear* up to a modulation index of 0.5 and that the higher-order sidebands are *almost* nonexistant at lower indices. **Fig. 2** shows the spectrum of a 100% a-m signal, a 50% a-m signal, a 1.0 order fm signal and a 0.5 order fm signal. Note that the 1.0 order fm signal succeeds in generating first-order sidebands comparable with those of the 100% a-m signal, but at the expense of producing 2nd and 3rd order sidebands of appreciable amplitude.

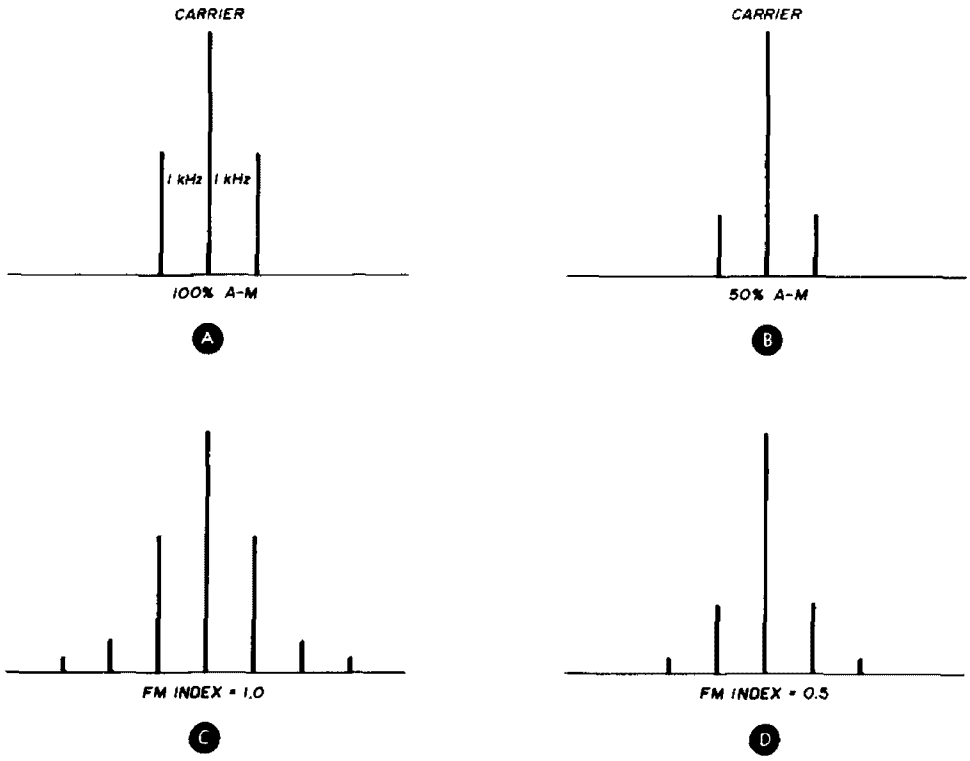


fig. 2. Frequency distribution for 50% and 100% amplitude modulation and fm signals with modulation indices of 0.5 and 1.0.

The 0.5 index fm signal gives a good approximation to a 50% a-m signal, with only small amplitude 2nd order sidebands. Since most amplitude modulated signal generators are only used at a-m percentages of about 50% (a standard measurement technique), you can use this 0.5 order fm signal to provide a simulated 50% a-m signal.

In fairness it must be mentioned that nbfm approximates low percentage a-m *only* in the frequency domain. The signal is still fm since there is no variation in amplitude at the modulation rate. That this is true is immediately obvious because the entire system is made of digital ICs which are in essence *limiters*; that is, amplitude is constrained to be either 1 or zero.

Since the amplitude does *not* vary, a diode detector will, strictly speaking, be unresponsive to nbfm. However, most receiver systems having diode detectors will respond to nbfm by the slope-detection mode, where the receiver selectivity curve provides a frequency-to-amplitude conversion.

the circuit

The circuit of the signal generator is shown in fig. 3. Note that half of an MC3029P line-driver NAND gate follows each of the two multivibrators in the MC4024P. The line-driver NAND gates provide isolation and the capability to drive 50-ohm lines with either 1-kHz or rf output. It must be remembered that the output of the generator is well over one volt peak-to-peak, even when terminated in 50 ohms, so an external attenuator is usually required.

The modulation frequency is determined by the parallel tuned circuit consisting of the 88-mH toroid, T1, and the 0.33μF capacitor across it. This is because the frequency of the 1-kHz oscillator is adjusted (by voltage control) to maximize the output at the test point. This occurs

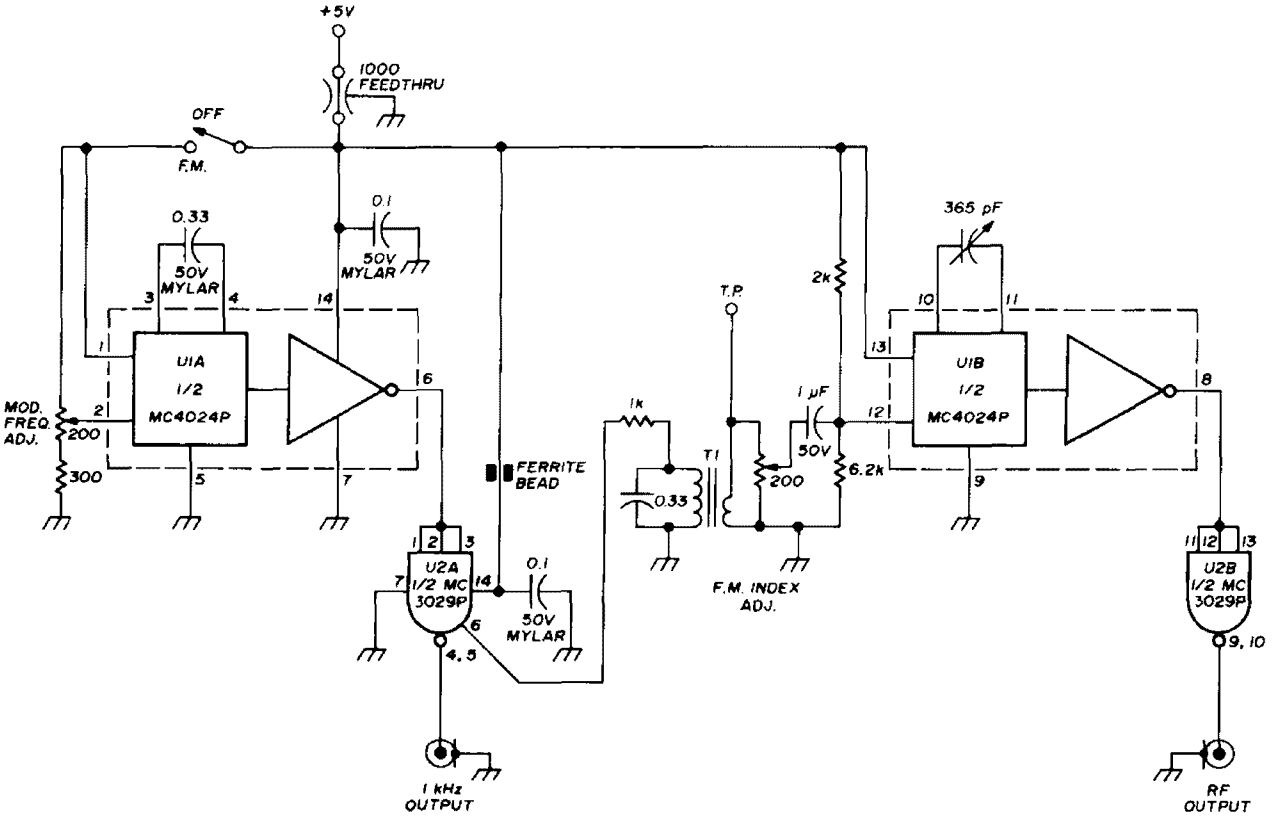


fig. 3. Circuit for the wide-range rf signal generator that covers from 600 kHz to 12 MHz. Integrated circuti U1A is the 1-kHz modulation oscillator; U1B is the rf oscillator. U2A and U2B are used as line drivers. Transformer T1 is an 88-mH toroid with a secondary consisting of 30 turns no. 28 enamelled wire wound over it.

when the 1-kHz oscillator frequency matches the resonant frequency of the 88-mH-0.33-μF parallel tuned circuit. The voltage observed at TP with a scope should be about 0.2 volts p-p. If the resultant frequency at maximum TP voltage isn't close enough to 1-kHz to suit you, a somewhat different value of C (nominally 0.33-μF) will have to be used.

A simple but well-regulated power supply is shown in fig. 4. It uses a standard 6.3-Vac filament transformer and a full-wave bridge rectifier. The regulation is accomplished by one of the newer three-terminal IC voltage regulators of Fairchild, National or Motorola. The common terminal of each of these regulators is the case, so a good thermal connection to the chassis (for heat dissipation) is also the electrical ground. The 0.22-μF capacitor at the input of the voltage regulator is important and should not be omitted. This capacitor should be placed between the input and common terminal of the regulator *immediately* adjacent to the IC to avoid any chance of instability.

The signal generator is built into a 7.5x4.7x3-inch Bud CU-347 cast aluminum box, which is mounted on an 11x7x2-inch aluminum chassis. A 7x12-inch panel is used to mount the Millen 10035 dial assembly. The broadcast variable capacitor is electrically "floating" inside the cast aluminum box and is mounted by screwing it to a Lucite plate, which in turn is mounted on 1/4-inch standoff spacers to the inside bottom of the box. The MC4024P and MC3029P ICs are socket-mounted upside-down on a 2-1/2x2-1/2-inch piece of double-sided copper-clad circuit board. The ICs themselves are not visible, but the socket terminals are conveniently exposed for wiring.

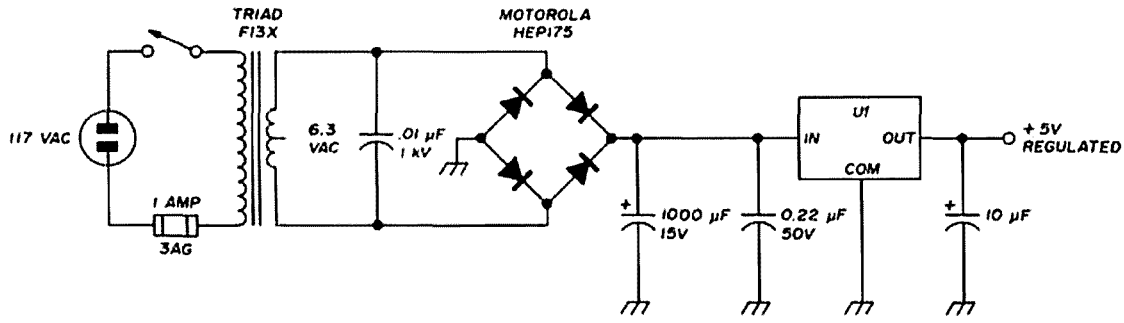
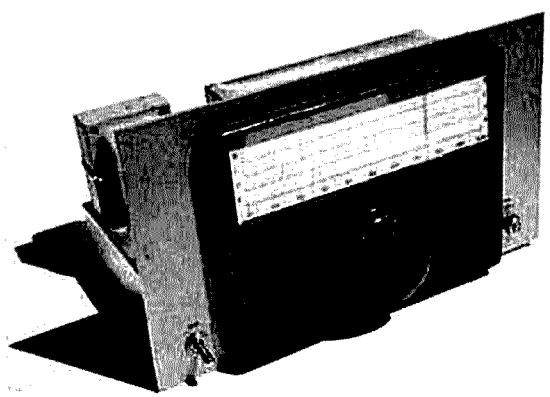


fig. 4. Regulated power supply for the wide-range signal generator. Three-terminal regulator U1 is a Fairchild μA7805, National Semiconductor LM309K or Motorola MLM309K.



The wideband signal generator is built into a Bud CU-47 enclosure. Dial mechanism is a Millen 10035.

The power supply circuitry is built in the underside of the 11x7x2-inch aluminum chassis. In this way all the parts of the power supply which have large 60-Hz signals on them are well isolated from the MC4024P — which has quite a high modulation sensitivity.

Since the broadcast variable capacitor I used has two 365-pF sections, only one of which is used, it would be possible to add frequency coverage down to 300-kHz by simply adding an spst switch. This was not done in the preliminary model because it was not mechanically convenient. However, such an addition should be considered when building a new version, since the 455- to 500-kHz region is quite useful for i-f alignment.

ham radio

two-stage cavity filter for two meters

Complete construction details for a highly selective resonant filter for 144 MHz

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Because of the popularity of the fm mode of communications, the amateur vhf bands are becoming much more active. With this activity comes the attendant equipment problems, which include interference to and from our landmobile service neighbors who, in some cases, use the same geographical location as the amateur station. Overloading the neighboring receiver, or being overloaded, are the most prominent problems. Spurious radiation is another nuisance.

Overloading manifests its presence by the sudden decrease in sensitivity of a receiver which has a signal forced into its input. The overloading signal does not have to be near the operating frequency of the overloaded receiver, but it will be strong enough to get into the frontend and cause the agc to cut down the overall gain of the receiver. Often, when this effect occurs, the operators will not be aware of it because the overload signal bears no intelligence. The reverse of this effect causes problems with the neighboring station in exactly the same manner. If any spurious radiation occurs at the same time you can be sure the amateur is the one to be roasted.

looking for the cure

To improve neighborhood relations with a technician who maintains equipment in the same building and uses the same antenna platform as I do, an investigation was completed which revealed the desensitization of several receivers. One receiver operated in the amateur two-meter fm band and another in the Land Mobile Service on an adjacent frequency allocation. The transmitters for each of these two services were at the 50-watt level.

A probe with a crystal detector was mounted on the tower, halfway between each antenna; meter indicators were located near each of the transmitter/receiver units so that observations of *on time* could be accurately known and used when comparison adjustments were being initiated. It was interesting to note that other services, 10 MHz away in frequency and located geographically on the other side of the hill, were detected and in several instances desensitized the commercial receiver. A plot of the input circuits for the rf amplifier and mixer for the amateur receiver was made. The input circuits and interstage coupling circuits are double tuned and critically coupled by the manufacturer, indicating that previous thought had been given to the matter of high-Q preselection.

A similar test was made on the commercial station receiver. The plot for the amateur receiver preselection circuits, fig. 1, is presented on a scale which clearly shows how the adjacent frequency transmitter could easily control its sensitivity through overloading. To eliminate this

problem it was obvious that further selectivity was required for the receiver front-end.

To accomplish this, a major circuit revision would be required. Further investigation of several of the commercial sets revealed that the same problem had been relieved satisfactorily by adding a coaxial filter to the antenna feedlines. These units were simple coaxial tanks, designed with a low coupling coefficient to maintain a high Q, and, therefore, improve selectivity.

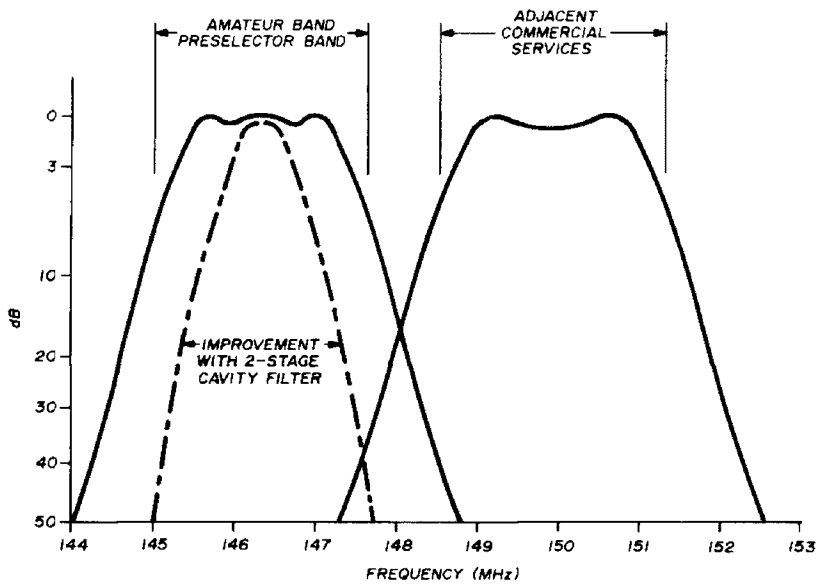


fig. 1. Typical selectivity curves of amateur two-meter receivers and adjacent Land Mobile equipment. Two meter selectivity is improved considerably by the addition of the two-stage cavity filter described here.

coaxial filter

A dual coaxial filter was designed for 145 MHz. The filter was built from plumbing house supplies because these parts are very readily available. Construction details are shown in fig. 2. A list of materials is included for 145 MHz which will assist the constructor in locating the required copper fittings (table 1).

Assembly of the multiple-cavity filter is simple. First, inspect two 1-1/2x3/4-inch reducing couplings to see that there are no dents on either perimeter. Next, carefully file smooth the lip found in the interior of the 3/4-inch entry. When filing try not to touch the smooth area of the 3/4-inch pipe wall on this fitting, just break down the step so that a piece of 3/4-inch pipe will slide through each fitting.

Lay out the holes to be drilled in the B section of 1-1/2-inch copper water pipe. The lengths of the pipe and hole locations can be determined from the chart accompanying fig. 3. Both holes should be concentric; one should be large enough to accomodate the round shoulder on the mounting flange of a SO-239 coaxial receptable. At a point 180° away from the connector hole, a second entry is required which is large enough to allow a piece of 3/8-inch copper water pipe to slip in to a tight fit.

At the base of each reducing coupler drill two holes with a number-28 drill. Slide the B section of 1-1/2-inch copper pipe into the reducing coupler. Align the reducing coupler so that the number-28 holes are parallel with the two large holes in B section. Sweat solder the two parts; use just enough heat to cause the copper to slightly change color. Use soldering paste. When the joint has been soldered, wash away the paste residue with hot water or a cleaning solvent. Try to make the joints as nearly watertight as possible.

The next step requires the addition of part H, a 1-1/16-inch disc which is soldered to the end of part J, a section of 3/4-inch copper water pipe, 17-inches long. Two sections should be prepared. The disc can be a large steel washer or can be cut from sheet copper. It is half of a capacitor used to foreshorten the cavity. It is also part of the tuning system.

Prepare a SO-239 coaxial receptacle by soldering a 3-inch length of number-14 wire to the center conductor terminal. Bend the wire at a right angle directly

where it exits the solder point on the connector, insert the end of the wire into the large hole provided for the connector on the B section of copper pipe, and feed the wire down so that it enters the drilled number-28 hole into the reducing coupling. Solder the SO-239 fitting into place. Likewise, solder the wire to the exterior base of the reducing coupling and trim

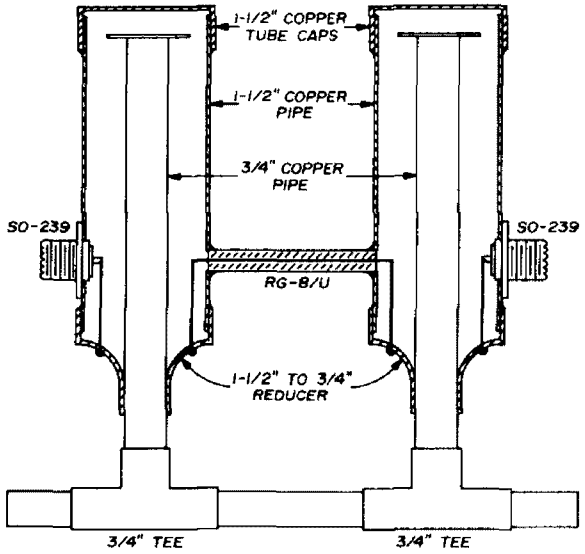


fig. 2. Two-stage cavity filter for two meters is built from common copper plumbing components.

off any excess wire and solder. Align the wire so it is parallel with the pipe wall. This completes the input coupling jack assembly.

output coupling

The next step provides the output coupling link. Prepare an 8-inch section of RG-8/U as follows: strip off the vinyl jacket and the shield braid. Measure 2-1/2-inches in from each end of the center dielectric and cut away the covering to expose the center conductor wire. Bend the wire 3/8 of an inch from the remaining covering, slide the end of the wire into the B section of the cavity through the 3/8-inch hole, and fish it down to the remaining number-28 hole in the reducing coupling. Prepare the second cavity in the same way, leaving out the last step.

Slide a 3/8-inch copper pipe, 3-1/8-inches long, over the coaxial-cable center dielectric and into the hole on the B section of the first cavity wall; carefully solder it in place. Bend the remaining end of the center conductor in the same manner as the opposite end, insert it into the 3/8-inch hole in the second cavity and into the number-28 hole located in the second reducing coupling. The end of the 3/8-inch pipe will now be fitted to the wall of the second cavity and soldered in place. The input/output and inter-cavity coupling are now complete.

The cavity assembly at this point is quite fragile and must be handled as such. Two 1-1/2-inch pipe stands must now be added to each cavity. One stand should be located at the top of each cavity, the other just above the 3/8-inch pipe containing the coaxial coupling element. Fasten the stands to a section of aluminum panel which will serve as a mounting for the filter. Level each pipe stand so that no strain is given to the inter-cavity coupling. If pipe stands are not available, a pipe clamp can be used.

table 1. List of materials required for the two-section two-meter cavity filter.

qty	description
2	1-1/2- to 3/4-inch copper reducing coupling (Mueller Streamline style WC400R, part number W1067)
2	3/4-inch drop ear tees, copper (Muller sytle 310, part number A1550)
2	1-1/2" copper tube caps (Mueller style WC415, part number 7013)
1	25-inch length of 1-1/2" copper water pipe, cut into two 12-1/2-inch lengths
1	34-inch length 3/4" copper water pipe, cut into two 17" lengths

The above material can be obtained from plumbing suppliers; part numbers are those of the Mueller Brass Company, a large plumbing parts manufacturer. Similar parts are available from Sears.

Now, take two 17-inch lengths of pipe with the washer or disc soldered in place and insert the open ends into the 3/4-inch opening of the reducing coupling. Slide each section down to point where the disc is 1/2-inch below the top of the B section.

Be sure to clean these two parts so

that they are very bright. Use fine steel-wool. Be sure the reducing section is also clean; these pieces will be soldered later and must make a very good connection. Place a tube cap on the open end of each cavity. Both contact points should be clean and bright since these parts also will be soldered. Push the cap down as far as it will go.

tune up

You are now ready to tune up the cavity. It is best to use the transmitter coupled through a vswr bridge to a 50-ohm load for tuneup as shown in the block diagram.

Since the vswr bridge will serve as a resonance indicator, it should be set to the *forward* position. Set the sensitivity control to minimum. When power is first applied to the filters a small indication will be observed. The end of each of the 3/4-inch pipes should be carefully moved in very small steps which will cause the vswr meter to indicate an increase in output. Adjust each pipe until there is no further increase in output level. The sensitivity of the bridge can be adjusted as required. The reflected vswr should be no worse than before the filter was inserted if you follow all the dimensions shown in fig. 2.

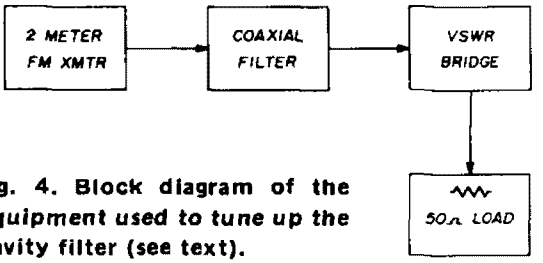


fig. 4. Block diagram of the equipment used to tune up the cavity filter (see text).

If you reverse the input and output connections, there should be no difference from the previous measurements. For this reason, it makes no difference which SO239 jack is used for the input or the output.

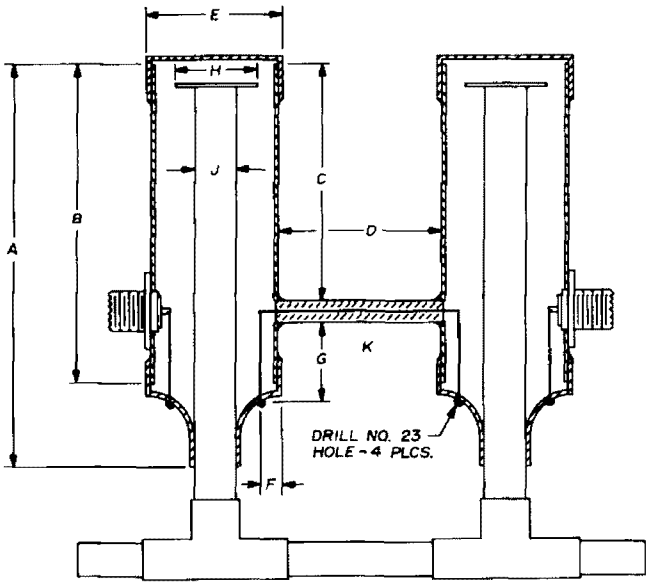
To determine the amount of signal loss through the filter simply connect the coax directly to the vswr bridge and note the level. Compare output power without the filter to the level of the output with the filter installed. If you want to determine the ratio of *on* frequency resonance to *off* frequency loss, simply switch the transmitter to the 144-MHz end of the band for the low-end ratio and to 149-MHz for the high end loss. Loss at the high and low ends of the band should be near 40 dB.

When the tuneup adjustments are complete, carefully solder the 3/4-inch pipe to the reducer entry. Solder the top cap in place. The filter is now complete and it can now be installed in the feedline of your transceiver. A set of dimensions for filters for other vhf bands is shown in fig. 3.

The improvement at my station has been worth all of the effort and at not too great a cost. I no longer have the desensitizing effect and my commercial neighbor now has a similar filter tuned up on his Land Mobile channel.

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fig. 3. Construction of the two-stage two-meter cavity filter. Dimensions for other vhf bands are shown in the attached chart.



	frequency MHz			
	50	144	220	440
A	41.0"	17.0"	7.0"	5.0"
B	38.0"	12.6"	5.0"	3.0"
C	31.6"	12.9"	5.0"	3.8"
D	3.0"	3.0"	3.0"	3.0"
E	4.5"	1.5"	3.0"	3.0"
F	1.4"	0.375"	1.0"	0.75"
G	6.0"	2.1"	1.4"	0.75"
H	3.0"	1.06"	2.75"	2.75"
J	0.75"	0.75"	0.75"	1.0"
K	----------see text----------			

three-terminal voltage-regulator ICs

Design and construction of regulated power supplies is simplified by the use of three-terminal voltage-regulator ICs

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The new Fairchild 7800-series three-terminal voltage-regulator ICs present some vastly new features not previously available to the amateur. For one thing, they provide a lot of regulation for very little money. However, the use of these regulators takes a new orientation, especially for those readers who have designed and/or built conventional regulated power supplies. In this article, I will cover some of the new aspects of using these IC regulators and will show several recommended circuits for different types of power supplies.

Many new integrated circuits no longer require the addition of several "generalized" circuit elements, such as transistors and diodes, to perform a specific function. Rather, these functional blocks are already combined, within the IC, to perform one very specific function.

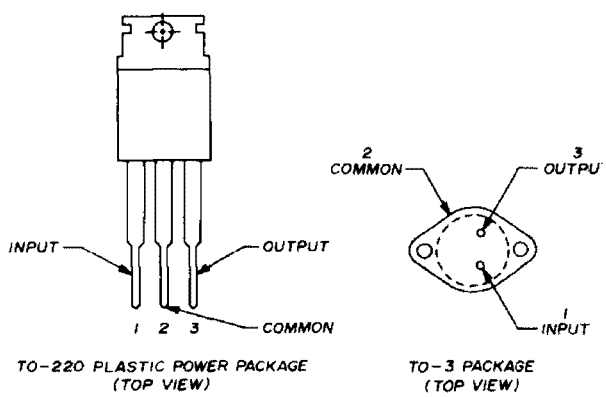


fig. 1. Electrical connections to the Fairchild 7800-series three-terminal voltage-regulator ICs.

The 7800 regulators are good examples of this new breed of IC. Each device in the series is preset to regulate a fixed output voltage. For example, the 7805 is a positive five-volt regulator. A complete list of the 7800 series and the respective preset voltages are shown in table 1.

The main advantage of a fixed-voltage regulator is the ease with which it can be used. Since the basic operation of the 7800 IC requires no external components, all you need is a power transformer, a bridge rectifier and a filter capacitor, and you have an instant power supply. You don't have to worry about choosing transistors, biasing them, and protecting the regulator against short circuits.

features

The 7800 voltage-regulator series features a preset voltage tolerance of ±5%, more than adequate for the vast majority of electronics experimenting. The tolerance means that the actual output voltage of an individual 7805 sample, for example, may be anywhere between 4.75 and 5.25 volts. However, the actual voltage regulation, once you have chosen a particular device, is 0.01% per volt, or 0.05% for the five-volt 7805. I doubt that most experimenters need better regulation than that!

Another valuable feature of the 7800 IC regulators is their built-in protective circuitry. The circuit guards against the three most common causes of power supply failures: excess output current, output short circuit and excess heat. The first two causes are listed separately because of the subtlety of a current overload — you may have your project hooked up properly, but are simply demanding a little too much current. The 7800 regulators compensate for these failure modes by internally limiting the output current that can be drawn from the device. In the case of a complete short circuit, only 750 mA, typically, can be drawn from the 7805.

The thermal shutdown protects the regulator from overheating. Additional safe-area compensation of the output transistor prevents the circuit from trying to dissipate too much power. Power capability is 15 watts. This means that you can draw 1 amp of current at 5 volts if the average unregulated input voltage is 20 volts or less, and if adequate heat sinking is provided.

The 7800 series ICs come in two case styles: a TO-220 plastic power transistor case, and a metal TO-3 case. The TO-3, having a lower case-to-ambient thermal resistance, is easier to heat sink, but it is more difficult to mount. Electrical connections are shown in fig. 1.

table 1. Low-cost three-terminal fixed-voltage IC regulators manufactured by Fairchild, Motorola, National Semiconductor and Silicon General.

Fairchild number	National number*	Motorola number	regulated voltage
μA7805	LM340T-5	MC7805	5 volts
μA7806	LM340T-6	MC7806	6 volts
μA7808	LM340T-8	MC7808	8 volts
μA7812	LM340T-12	MC7812	12 volts
μA7815	LM340T-15	MC7815	15 volts
μA7818	LM340T-18	MC7818	18 volts
μA7824	LM340T-24	MC7824	24 volts

*The letter T designates the TO-220 package; for the metal TO-3 package, substitute the letter K. Motorola devices are in a metal TO-220 package.

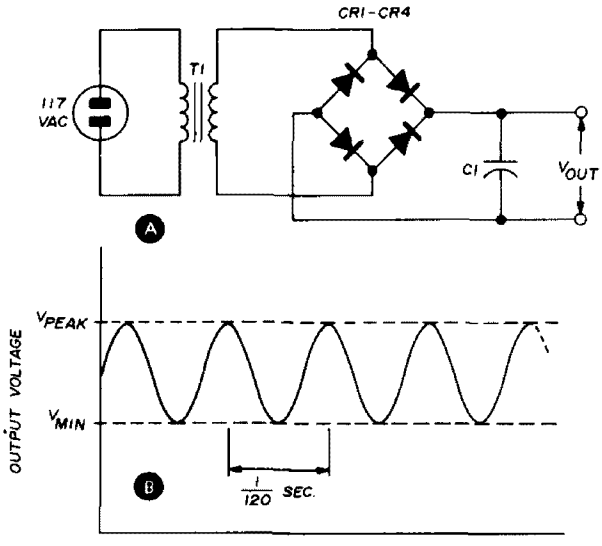


fig. 2. Basic unregulated power supply and typical 120-Hz ripple component.

unregulated supply

The unregulated power supply is a

basic element for the properly operating regulated supply. An example of an unregulated power supply is shown in fig. 2A. The line voltage is stepped down by transformer T1, rectified by the diode bridge and filtered by the output capacitor. With no load, the output voltage is equal to the peak transformer output

design the unregulated supply. For all the regulated circuits discussed here we will assume a properly designed unregulated supply which can provide 1 ampere of current without allowing V_{min} to drop below the sum of the desired output voltage *and* the regulator maximum voltage drop.

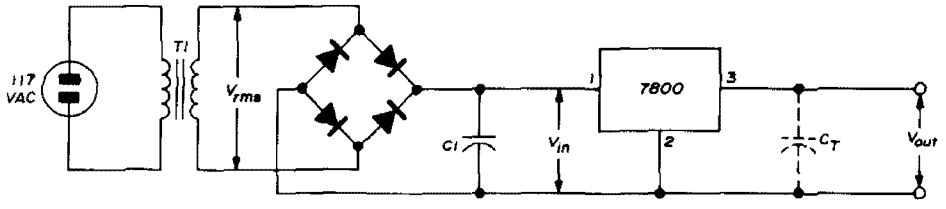


fig. 3. Basic positive voltage-regulated power supply using a 7800 series IC. Capacitor C_T may be added to improve transient response.

voltage (1.4 times the rms voltage), less twice the diode forward voltage drop. As current is drawn, however, the voltage decreases momentarily between the charging peaks of the bridge. This creates the ripple shown in fig. 2B.

The average output voltage is between V_{peak} and V_{min}. The greater the load, the more capacitor C1 discharges between charging pulses, and the lower V_{min} becomes, thus accentuating the ripple. For the same load current, the ripple decreases with an increase in the size of C1, up to the point where the bridge can

designing the unregulated supply

Although the discussion which follows pertains to the 5-volt 7805, a similar design approach is used with other members of this voltage regulator family. The 7800 series of ICs requires a minimum of 2 volts input-output differential for proper regulation. This means that V_{in}, in fig. 3, must be at least 2 volts higher than V_{out}. Since the preset voltage of the output is ±5%, the worst case is 1.05 times the rated voltage, or

$$V_{out\ (max)} = -1.05\ V_{out}$$

table 2. Operating parameters for an unregulated supply designed for use with a 7805 5-volt IC regulator.

Desired operating voltage	V_{out}	5 volts dc
5% tolerance (1.05 V_{out})	$V_{out\ (max)}$	5.25 volts dc
Differential (+ 2 volts)	$V_{in\ (min)}$	7.25 volts dc
Ripple allowance (1.1 $V_{in\ (min)}$	V_{peak}	7.98 volts dc
Diode drop (+ 1 volt)	V_{peak}	8.98 volts dc
10% line variation (1.1 V_{peak})	V_{peak}	9.88 volts dc
Transformer output (V_{peak}/1.4)	V_{rms}	7.06 volts rms

no longer recharge C1 fast enough. One way to eliminate the output ripple is by regulating the output voltage. However, you must never drain so much current as to allow V_{min} to dip below the desired regulated output voltage.

Thus, the first step in building a regulated power supply is to properly

Since there must be at least a 2-volt differential across the 7800

$$V_{in\ (min)} = V_{out\ (max)} + 2$$

With a 10% ripple, at full current, the peak value of V_{in} should be 1.1 $V_{in\ (min)}$. Since the transformer current must pass through two of the bridge diodes

during any charge pulse, two diode drops (approximately 0.5 volts) must be added, 1.1 $V_{in\ (min)}$ + 1.0 volt. This is the peak output voltage required from the transformer. If you allow for a 10% variation in the line voltage (105.3 to 128.7 volts for a nominal 117-volt line), you require an extra 10% for the transformer output

regulator circuits

The basic hook-up for the 7800 voltage-regulator ICs is refreshingly straightforward, as shown in fig. 3. The only embellishment is the optional transient suppression capacitor, C_T, at the regulator output. This capacitor, typically 10- to 50-μF, will improve transient and

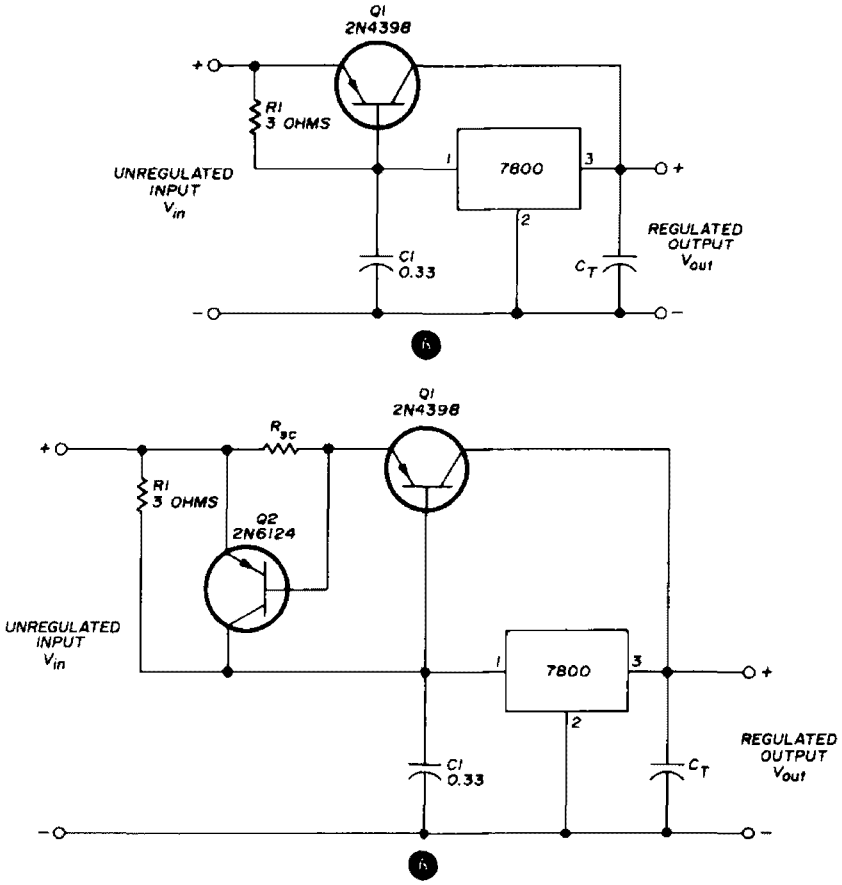


fig. 4. High current voltage regulator circuit using external power transistors. Circuit in (B) includes short-circuit protection for the power transistor (see text).

voltage. These considerations are summarized for a 5-volt supply in table 2. In the worst case, the transformer must be able to produce at least 7 volts rms to operate properly.

The only thing that remains is to choose capacitor C1. The capacitor must have sufficient capacitance to prevent its voltage dropping below 7.25 volts when 1 amp is being drawn. A few quick calculations will show that a capacitance of 12,000 μF will meet this condition. Of course, if the actual transformer voltage chosen is greater than the 7.06-volt minimum, a smaller capacitor will do fine, since more ripple can be tolerated.

high-frequency response, but at the cost of increasing output impedance at frequencies below 1 kHz. Additionally, if a battery is used for the unregulated supply, an input bypass capacitor of at least 0.22 μF should be attached across the input to the 7800 (pin 1 to ground).

To increase the current capacity of the 7800s ICs, you may wish to add a pnp series pass transistor as shown in fig. 4A. In this application, the pass transistor handles most of the supply current. The 2N4398 transistor shown has a maximum collector current of 30 amperes. The 7800 regulator IC holds the output voltage constant by varying the bias on the

base of the pass transistor. As the load current increases, the output voltage drops slightly, causing the 7800 to draw more current. This increases the base current of transistor Q1, which brings the output voltage back up by supplying

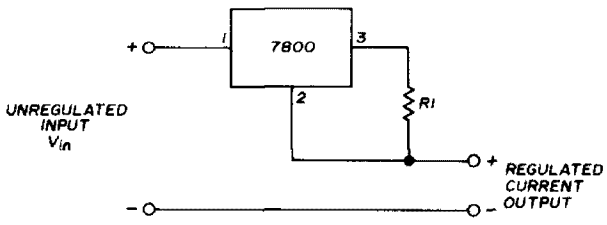


fig. 5. Circuit for using the 7800-series voltage-regulator ICs as current-mode regulators.

more current to the load. Thus, the output current is supplied by both the 7800 IC and the pass transistor.

If you are concerned about a short circuit burning out the power transistor, you can insert a protection circuit consisting of transistor Q2 and resistor R_{sc}, as shown in fig. **4B**. The 7800 IC protects itself, but has no feedback feature to protect external elements. The series resistor, R_{sc}, may be set for the particular value of current you wish to limit. It should be a small value resistance capable of handling the power through it. Transistor Q2 may be any moderate gain npn transistor which can handle the short-circuit current of the 7800.

The use of this series pass transistor allows you to quickly build a fixed regulated voltage supply for almost any application you may have. In fact, if the protection circuit of fig. **4B** is used, R_{SC} may be made variable, so as to provide exactly the extent of current limiting needed to protect the load circuit.

Some applications, such as battery charging, require a constant load current, rather than a constant voltage. Fig. 5 shows the connections to a 7800 for current regulation. The 7800 tries to maintain a constant voltage across R1. In the case of the 7805, the voltage is 5 volts. Obviously, the 7800 will regulate the amount of current through R1 necessary to maintain this voltage. As this supplied current also passes through the load resistance, its current is likewise regulated. In the case of the 7805, five volts across a load of 100 ohms would produce a load current of 50 mA – the charging current for a 500 mA-hr nicad battery. Within the limits of the unregulated input voltage, the 50 mA will be supplied equally well to a simple battery, or a whole stack of them, and the charging current will not change as the cell voltage goes up during the charge cycle.

summary

In conclusion, you should find the 7800 IC voltage regulators to be an invaluable addition to your electronics repertoire. It will free you from the frequent and routine task of building regulated power supplies. Also, you will find yourself using regulated supplies more often because of the simplicity and low cost ($2.20 in single quantities) of the 7800 voltage-regulator ICs.

ham radio



"Remember your application for renewal of your ham license that I was supposed to have mailed four months ago?"

low-voltage audio agc amplifier

Description of a wide range audio agc system that operates from a 1.5-volt flashlight battery

Courtney Hall, WA5SNZ, 7716 La Verdura Drive, Dallas, Texas 75240

The circuit described here operates from a single penlight flashlight battery with a current drain of 0.5 mA, nominal. It should be ideal as a self-contained unit which can be connected in a microphone cable. The agc control element is a transistor used in the inverted connection to obtain better performance. Those readers who are unfamiliar with audio agc theory and applications are referred to a previous article.[1]

circuit

Fig. 1 shows a schematic of the agc amplifier. Transistors Q2, Q3 and Q4 form a 70-dB voltage amplifier; Q5 is the detector, and Q6 is an emitter follower required to drive the control transistor, Q1.

Resistor R1 and transistor Q1 form a voltage divider which attenuates the signal, as needed, to hold the amplifier output constant. When the input signal is 50 μV or less, the detector has zero output, and Q1 is turned off. As the signal increases, the detector feeds dc to the base of Q1 causing its collector-emitter resistance to decrease; this decreases the signal input to Q2.

With a power supply voltage of only 1.5 volt, the detector must be able to operate from a relatively small peak-to-peak ac voltage, or clipping will occur. The detector shown requires only 0.62 volt peak-to-peak input.

Notice that the control transistor, Q1,

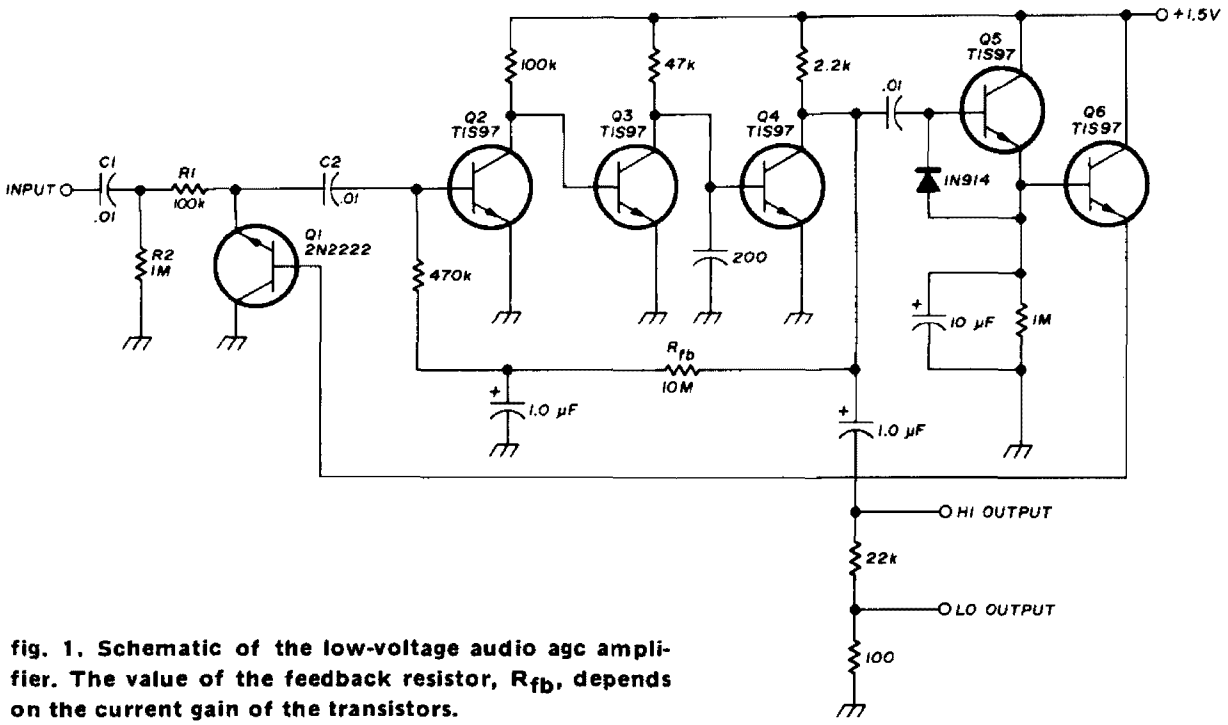


fig. 1. Schematic of the low-voltage audio agc amplifier. The value of the feedback resistor, R_{fb}, depends on the current gain of the transistors.

is shown in the inverted connection as is used with chopper transistors; the transistor is turned on by current flow through the base-collector junction. Current gain is very low in this configuration, being on the order of 1.0, but collector-emitter resistance vs base current is about the same as with the normal connection.

The inverted connection is preferred because it produces a very low offset voltage; offset voltage is the dc voltage appearing between the collector and emitter when the base is biased on with no collector power supply. Fig. 2 shows how offset voltage is measured. Offset voltage in the normal connection can be 50 mV or more, but it is about 1 mV or less in the inverted connection.

Why is low offset voltage important? When the input signal to the agc amplifier suddenly increases from a very low value to a large value, the detector turns the control transistor on rapidly. If the control element produces a significant dc voltage across its terminals, a transient voltage spike is coupled to the amplifier input. This spike bears no relation to the signal amplitude and can drive the amplifier into hard saturation.

Suppose a 30-mV dc level suddenly appeared across the control transistor. The 70-dB gain of the amplifier would try to amplify the transient to a level of about 100 volts peak. Naturally, the amplifier cannot do this, so it saturates, upsetting the quiescent bias conditions. Recovery time from this strong transient may be one second or more. Under such conditions it is virtually impossible to achieve fast attack times.

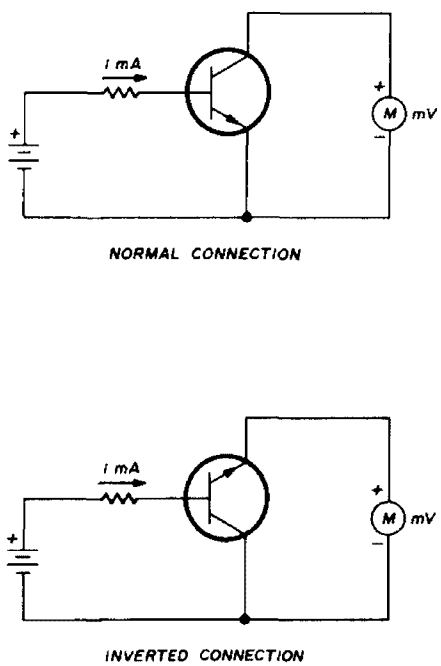


fig. 2. Procedure for measuring the offset voltage in the normal and inverted transistor connection with a millivolt meter connected between the collector and emitter.

Several transistor types were tried in the control transistor socket, but the 2N2222 was the only one that performed well. A transistor designed specifically for chopper applications should work best of all.

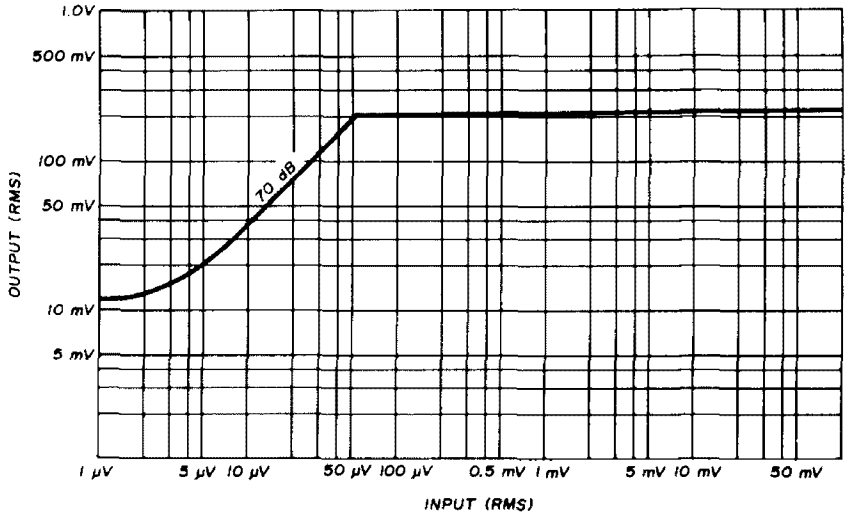


fig. 3. Input/output voltage characteristics of the low-voltage audio agc amplifier.

It is important not to omit the 1-megohm resistor, R2, because it prevents the buildup of dc voltage on the emitter of Q1 due to the capacitive voltage divider formed by C1 and C2.

Two outputs are shown in fig. 1; one is the full output of the amplifier, and the other attenuates the output to a level of about 1 mV rms, maximum. The attenuated output should be used when feeding the microphone input of other equipment so it will not be overdriven.

The value of the 10-megohm resistor, R_{fb}, depends on the current gain of the amplifier transistors, and its value may need to be adjusted. The value should be set so that the dc voltage on the collector of Q4 is about 0.8 volt.

operating characteristics

Fig. 3 shows the input vs output voltage curve of the agc amplifier. Although the maximum input voltage shown is 100 mV, inputs up to 1.0 volt rms may be applied without significant distortion of the output. The 3-dB bandwidth of the amplifier is approximately 100 Hz to 8 kHz; attack time is less than 50 milliseconds, and release time is about 2 seconds.

conclusion

The results achieved with this circuit show that good performance can be obtained from a 1.5-volt agc amplifier, and that suitable transistors used in the inverted connection for the control element offer improved agc characteristics.

reference

1. C. Hall, WA5SNZ, "Audio AGC Principles and Practice," *ham radio,* June, 1971, page 28.

ham radio



"The last time we went out to eat was because of power failure!"

bandpass filter design

How to design image parameter bandpass filters using the lowpass/bandpass analogy

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Very few amateurs attempt to design bandpass or band-stop filters, possibly because of the complex mathematical formulas involved. Most amateurs seem to be unaware of the analogy between lowpass and bandpass filters which can considerably reduce the amount of labor involved in the design of bandpass filters. This analogy has long been known in professional circles[1] but apparently hasn't been described in the amateur literature.

The principal purpose of this article is to describe the lowpass/bandpass (LPBP) analogy. A secondary purpose is to provide some tips on filter design that I have found useful in both amateur and professional applications; specifically, how to design filters using only a reactance slide rule or resonance calculator such as the Shure Brothers or Allied Radio reactance slide rule.

It has been my experience that the biggest difficulty most amateurs find in filter design is simply one of making arithmetical mistakes in computing the parameters — slide-rule errors, plain and simple. When you get the parameters right, the filters usually work! Therefore, anything that will mechanize the calculations or serve to check the calculations is a big help in building filters that work the first time.

The LPBP analogy has the further advantage of giving the filter designer a much better physical insight into the

practical constraints imposed on bandpass filters than does a set of cold mathematical equations.

image parameter filters

The type of filters to be discussed are known as "image parameter" filters. Historically, this type of filter was the first to be developed and is entirely suitable for many amateur and professional applications. For completeness I will begin with a brief discussion of image parameter filter design.

A low-pass filter consists of a series of sections as shown in the upper portion of fig. 1. As long as all sections are designed for the same impedance level, as many sections as are necessary may be connected in series to give the desired frequency response characteristics.

For convenience, filters are designed on a section basis, the basic section being known as a "prototype" section. After the prototype section has been specified, many different variations are possible, depending on the particular application. The sections are broken out of the composite filter in either of two ways: a mid-shunt (or pi-section) and a mid-series (or tee-section). Notice in fig. 1 that the value of shunt capacitor in the pi-section is one-half that of the composite filter; when two pi-sections are connected together the value of shunt capacitance adds up to the full value of shunt capacitance.

A similar statement can be made for the series inductance of the tee-section. The equations to determine the full value of both the series inductance and shunt capacitance in terms of the cutoff frequency and impedance levels are also given in fig. 1. These values are known as the prototype values; one-half of either the inductance or capacitance value must be used in the actual prototype section, depending on whether a pi- or tee-section is used. So as not to complicate the discussion, in the material that follows I will stick with the pi-section.

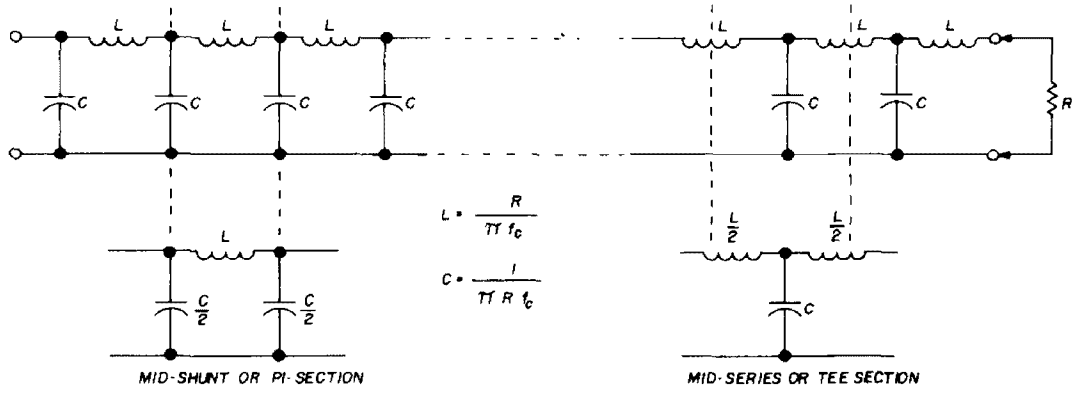


fig. 1. Basic filter sections with prototype design equations. R is the load resistance, f_c is the cutoff frequency.

First, take the equations for the prototype L and C as given in fig. 1 and put these into the resonant frequency formula to obtain

$$f = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi\sqrt{\frac{R}{\pi f_c} \cdot \frac{1}{\pi R f_c}}} = \frac{f_c}{2} \quad (1)$$

You can see that the prototype L and C values should resonate at one-half the specified cut-off frequency. This provides the designer with a means of using a reactance slide rule to compute (or check) the prototype parameters.

At this point a numerical example is

useful: Design an audio filter with a 3-kHz cutoff frequency working into a 2000-ohm load resistance. One-half of 3000 is 1500 Hz. Set the reactance rule to 1500 Hz and opposite 2000 ohms read approximately 0.22 henry and 0.054 μF. The prototype section appears as in fig. 2.

It is important to realize that the actual resonant frequency of the chosen inductance and capacitance is critical if the filter is to operate as desired. Although some liberties can be taken with the impedance values, the resonant frequencies should be as exact as possible. Therefore, care should be taken to insure that values of inductance and capacitance will resonate at the design frequency, even if the resulting impedance value is not exact. For this reason I suggest that the prototype values of L and C be calculated with a conventional *slide* rule or calculator, using the formulas, and checked with the reactance slide rule.

At this point a second check point is convenient. I previously pointed out that the resonant frequency of L and C must be one-half the desired cutoff frequency

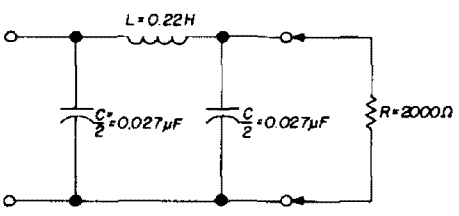


fig. 2. Basic prototype filter section with a 3000-Hz cutoff frequency.

of the filter; because the value of capacitance actually used in the prototype section is one-half that calculated above, the resonant frequency of inductance and capacitance *actually used* in the prototype section will be $1/\sqrt{2}$ or 70.7-percent of the desired cutoff frequency. This is easily verified with the reactance slide rule. In this case 0.027 μF and 0.22 H resonate at 2100 Hz which is 70.7 percent of 3000 Hz; hence, the values check and the complete filter design could proceed with confidence.

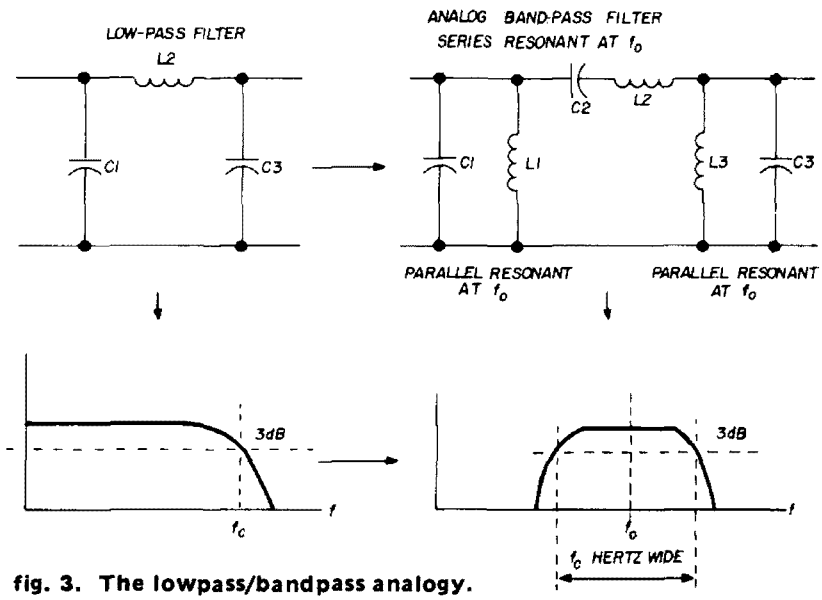


fig. 3. The lowpass/bandpass analogy.

lowpass/bandpass analogy

The LPBP analogy states that to transform a lowpass filter with a cutoff frequency f_c to a bandpass filter with a total bandwidth f_c and a center frequency f_o, it is only necessary to add inductance in parallel with each shunt capacitor so that the combination is resonant at f_o, and to add a capacitor in series with the original series inductance; this combination also resonates at f_o. This is shown in **fig. 3** along with the appropriate response curves.

If there is any confusion at this point, a numerical example should help clear things up. Let's design a bandpass filter centered at 455 kHz with a 15-kHz bandpass such as would be suitable for an fm receiver. The load impedance will be assumed to be 5000 ohms.

First, calculate the inductance and capacitance values for a lowpass filter having a cutoff frequency of 15 kHz:

$$C = \frac{1}{\pi R f_c} = \frac{1}{\pi(5000)(15000)} = 0.00425\ \mu F$$

$$L = \frac{R}{\pi f_c} = \frac{5000}{\pi(15000)} = 0.106 \text{ henry}$$

The basic lowpass filter is shown in **fig. 4.**

A quick check of these values with a resonance slide rule shows that the prototype values, C = 0.00425 μF and L = 0.106 H, resonate at 7500 Hz which is one-half 15 kHz, while the values of inductance and capacitance actually used, C = 0.00212 μF and L = 0.106 H, resonate at 10.6 kHz which is 70.7 percent of 15

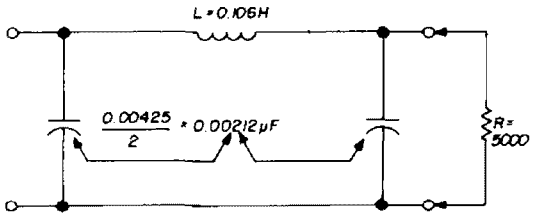


fig. 4. The first step in the lowpass/bandpass analogy — development of the prototype lowpass filter.

kHz. Therefore, we are on a firm foundation and can proceed with confidence.

The first step in transforming a lowpass section into a bandpass section is to connect an inductance in parallel with the shunt capacitor to resonate at 455 kHz, the bandpass center frequency. This is easily done by using a resonance calculator to obtain a shunt inductance of 58 μH. The second step is to connect a capacitor in series with the series inductance. For this example the capacitance should be 0.001155 μF. The completed bandpass filter is shown in **fig. 5.**

The difference between using the LPBP analog and calculating the individual component values may be compared by considering the bandpass circuit equations shown in **fig. 6.** The interested reader may verify the component values obtained using the LPBP analog by actually solving the equations given in **fig. 6.** The ease of using the LPBP analog will be obvious.

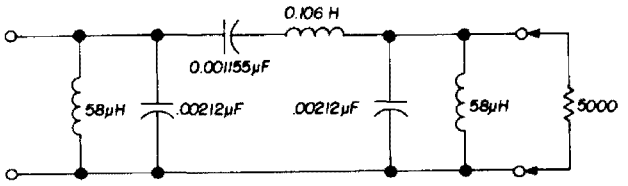


fig. 5. Analog bandpass filter section.

practical filters

The lowpass/bandpass analog also has the advantage that it can give the filter designer a much better feel for the physical realities involved. As an example of this I will attempt to design an exceptionally narrow bandpass filter in terms of center frequency, a 1-MHz wide filter with a center frequency of 50.5 MHz at 50 ohms.

Using the concepts described above, first calculate a lowpass filter with a 1-MHz cutoff frequency at 50 ohms. For a pi-section the inductance will have a reactance of 50 ohms at 500 kHz, or 15.9 μH. The capacitance *actually used* must resonate with this inductance at 70.7 percent of 1 MHz or 707 kHz. This is 0.00319 μF. So far, so good; the lowpass filter is shown in **fig. 7.**

The next step is to transform the lowpass filter into a bandpass filter. Now comes the joker which is easily seen from the LPBP analogy. To make this transformation it is necessary to resonate the 3190-pF shunt capacitor to 50.5 MHz. Now, 3190 pF is a lot of capacitance at 50.5 MHz; it is impractical to resonate

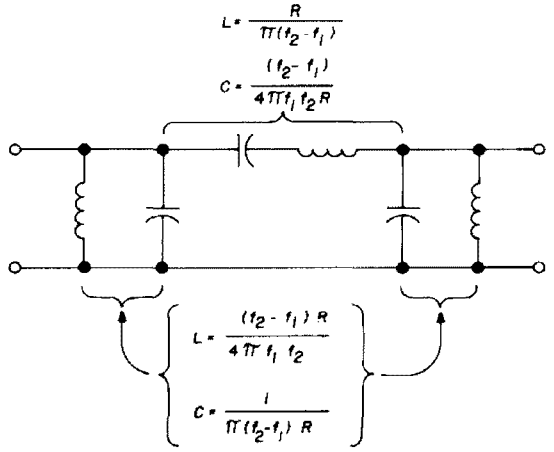


fig. 6. Design equations for the bandpass filter section. Frequency f1 is the low-frequency cutoff, while f2 is the high-frequency cutoff.

this much capacitance with any reasonable inductance. Resonating the series inductance to 50.5 MHz presents no serious problem and requires about 6 pF.

Thus, you can see from the LPBP analogy why a filter of this type is not practical in this application; the bandwidth is too narrow in terms of the center frequency. Before giving up, however, let's try a tactic frequently used by filter designers when one or more of the components turns out to be an impractical value — change the impedance level.

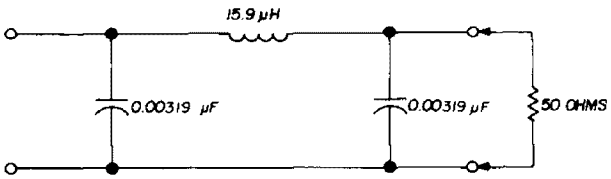


fig. 7. First step in the development of a 1-MHz wide bandpass filter centered at 50.5 MHz.

If the impedance level is increased to 5000 ohms, a factor of one hundred, the shunt capacitors decrease in value to 31.9 pF and the shunt inductances become 0.31 μH, both of which are at least in the ballpark of being practical. The value of series inductance increases, however, to 1590 μH, requiring only about 0.06 pF to resonate at 50.5 MHz. This is an impractically small value.

Therefore, it appears that a filter centered at 50.5 MHz with only a 1-MHz bandwidth is impractical at any impedance level, and other types of filters, such as coupled tuned circuits, must be used to obtain the desired selectivity. These filters, however, are beyond the scope of this article.

The preceding example shows how the LPBP analogy gives the filter designer a much better feel for the practical problems involved than does a purely mechanical application of the design formulas. The 455-kHz bandpass filter with a 15-kHz bandwidth had a lower ratio of center frequency to bandwidth, and also operated at a considerably lower frequency so that component parameters were much more realistic.

summary

After reading the above material you may ask, "Does the LPBP analogy work in the opposite direction; i.e., is there a bandpass to lowpass analogy?" The answer is, "Yes, provided all tuned circuits are tuned to the same center frequency."

It should also be noted that the LPBP analogy described here gives only one particular class of bandpass filter. There are many other types of bandpass filters, the most notable of which is probably a series of tuned circuits coupled by means of capacitance or inductance. The design of coupled tuned circuits is a subject in itself.

Although I have used an image parameter designed lowpass filter as the starting point in this article, the analogy applies equally well to lowpass filters designed on a Butterworth or Tchebyscheff basis.

reference

1. Vernon D. Landon, "The Band-Pass/Low-Pass Analogy," Proceedings of the IRE, December, 1936, page 1582.

ham radio



"Since you're a married man, the first thing you'll need to set up an amateur station is a very understanding wife."

introduction to the digital mixer

How to use the D-type flip-flop IC as a frequency mixer

Gerd H. Schrick, WB8IFM, 4741 Harlou Drive, Dayton, Ohio 45432

Basically, a mixer can be thought of as a switch operating at one frequency which will or will not pass a signal at another frequency. A signal interrupted in this manner generates a combination of various new frequencies with the difference- or intermediate-frequency as the desirable output. Normally, this desired frequency is filtered out for further processing.

In practice the requirements for a mixer are a nonlinear switching device and a large injection signal. These two requirements complement one another to a certain extent since a very nonlinear device requires less oscillator injection, while more injection is required with less mixer nonlinearity.

When working with digital circuitry you are dealing with two voltage levels; therefore, all signals must be square waves or close to it. This means that if you use a smooth sinusoidal signal, it must be converted to a square wave with a circuit such as a Schmidt trigger. If the sinewave is large enough, a simple diode clipper will do the job.

The digital equivalent of a simple frequency mixer is a gate with two inputs such as the 7400 IC. The output contains all the frequencies because the gate does not have a memory and follows momentary changes of either frequency; the desired output frequency must be filtered out.

If you use an edge-triggered D-type (delay) flip-flop such as the 7474 as a mixer, the leading edge of the square-wave oscillator pulse transfers the input signal to the output, and the output remains at this new level until the next oscillator pulse samples the input signal as shown in fig. 1. When the oscillator pulse is out of step with the input signal it turns off the output. Thus, the output is a square wave at the intermediate frequency which needs no filtering, except possibly to remove the odd-order harmonics.

There are several D-type flip-flops which can be used for this application. The common TTL 7474 can be used up to about 25 MHz. The high-frequency version of this IC, the 74H74, is usable up

to 43 MHz. The Schottky TTL version, the 74S74, can be operated to 100 MHz. For even higher frequency use, Motorola has introduced the new MECL MC12000 digital mixer, which is a D-type flip-flop which can be used up to 250 MHz. The MC12000 has built-in logic converters so it can be interfaced directly with slower TTL 74-series ICs.

Since this mixing scheme is basically a sampling technique, the ratio of the two frequencies to be mixed cannot exceed 2:1. That is, if the input frequency is lower than the oscillator frequency, it cannot be less than half; if the input frequency is higher, it cannot be higher than twice the oscillator frequency.

A typical digital mixer circuit using TTL ICs is shown in fig. 2. In this circuit a 7400 TTL gate is operated as a crystal oscillator. The other two gates of the 7400 are used as input buffers to the mixer, a 7474 D-type flip-flop. In this circuit the rf input signal must be lower than the crystal frequency, and the i-f signal must be less than half the crystal oscillator frequency.

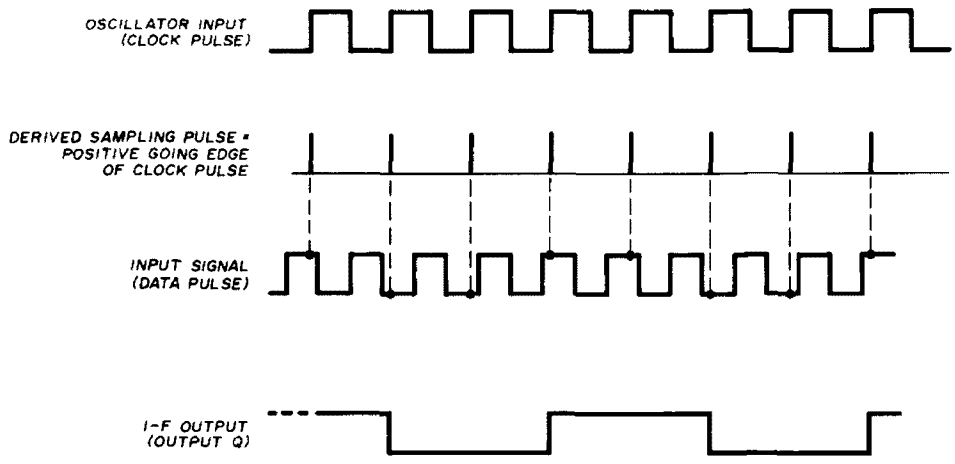


fig. 1. Waveforms in the digital mixer. Information from the input signal (data pulse) is transferred to the output by the positive going edge of the oscillator pulse (clock pulse). When the oscillator input is at either the high or low level, the signal input has no effect.

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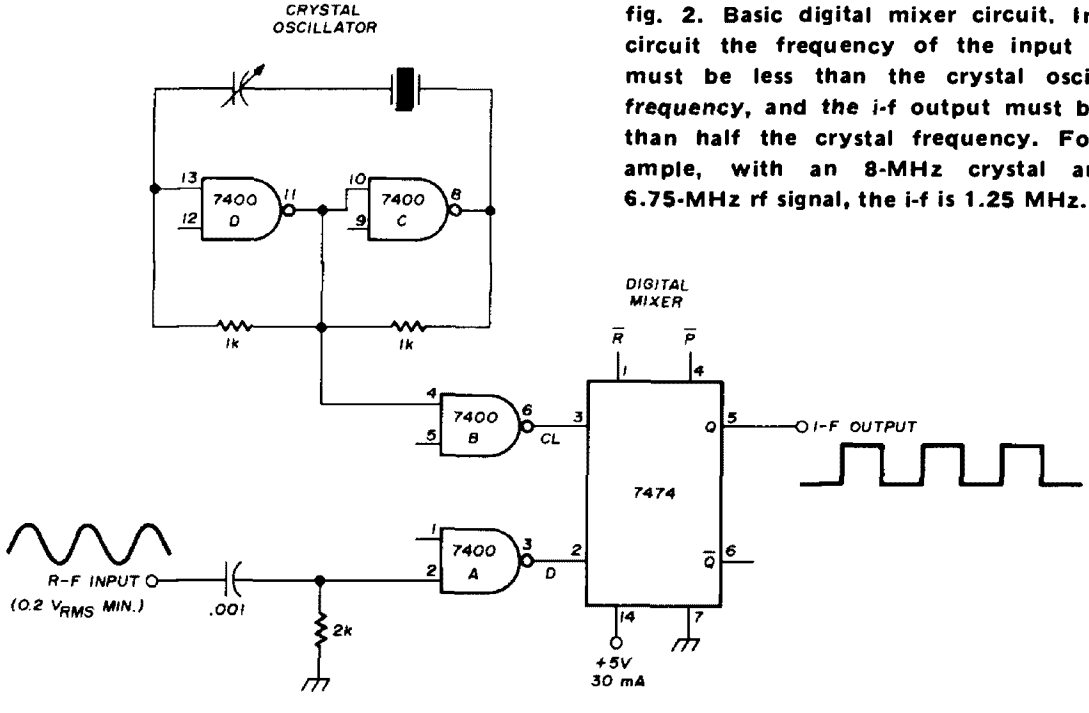


fig. 2. Basic digital mixer circuit. In this circuit the frequency of the input signal must be less than the crystal oscillator frequency, and the i-f output must be less than half the crystal frequency. For example, with an 8-MHz crystal and a 6.75-MHz rf signal, the i-f is 1.25 MHz.

narrowband modifications for the Regency HR-2 series of vhf-fm transceivers

How to install the Regency narrowband kit in the popular HR-2 series of two-meter fm transceivers

Paul Dobosz, WA8TMP, 2500 S. Hickory Ridge Road, Milford, Michigan 48042

The extremely popular HR-2 series of two-meter fm equipment introduced by Regency in 1970 has become an amateur favorite. At the time of introduction the desirability for an extremely selective narrowband receiver was not evident. This prompted Regency to build the HR-2 as a wideband unit.

As an increasing number of repeaters go into operation the wideband fm transceiver is plagued by annoying adjacent-channel interference. In many metropolitan areas all the repeaters are operating narrowband, in and out (deviation of ±5 kHz). With the large number of repeaters and growing popularity of two-meter fm, narrowbanding to conserve operating space is imperative.

The entire Regency family of HR-2 transceivers (HR-2, 2A, 2S and 2MS) can be easily modified for narrowband operation. Narrowbanding the transmitter is accomplished simply by setting the deviation control for a peak deviation of ±5 kHz. Narrowbanding the receiver requires some new parts and a few simple adjustments.

Before installing this modification in my receiver, I was plagued with adjacent-

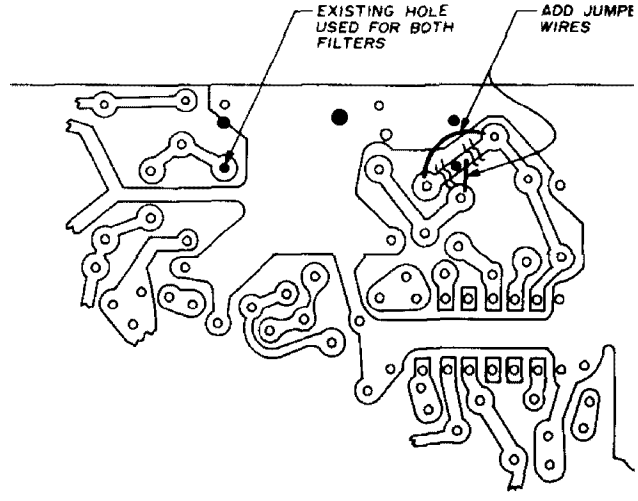


fig. 1. Bottom view of the HR-2 series receiver i-f circuit board, showing modifications required before installing the Regency MA-46 modification kit. One circuit trace must be cut in two places. The black dots indicate new holes for the new, narrow-band ceramic filter supplied with the kit of parts. A drilling template is shown in fig. 2.

channel interference on 146.76 MHz from the local repeater on 146.79 MHz. It was impossible to copy anything on 146.76 when the local repeater was transmitting. This repeater has an effective radiated power of more than 60 watts and is located less than a mile away from my station. With the circuit modification

MA-46 Narrow Band Filter (70-dB) Modification, the i-f board is identical to the one Regency uses in their FCC type-accepted marine and fm business-band equipment.

If you are the owner of a HR-2A, 2S or 2MS the instruction sheet provides all the information you need to install the

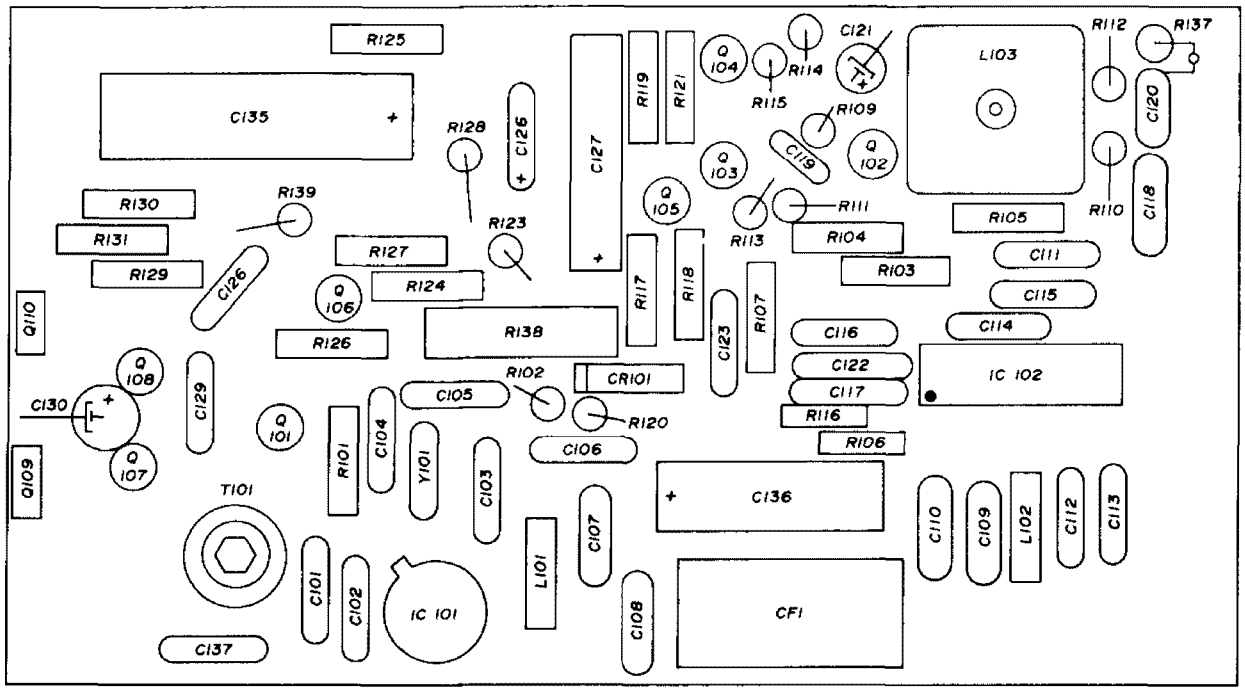


fig. 3. Parts layout of the Regency HR-2 series receiver i-f board. R133, R134, R135, R136, C134 and CR102 are located on the bottom of the board.

it is impossible to tell if the local repeater is operating or not while listening on 146.76 — even when the squelch is turned off.

modification kit

The basic kit of parts necessary to modify the HR-2 series transceivers is available from Regency. The kit consists of a new higher quality ceramic filter with an extremely steep selectivity curve that virtually eliminates adjacent-channel interference, two shielded coils to replace unshielded ones originally supplied with the rig, three capacitors to adapt the i-f circuit to the new ceramic filter, and two resistors to change the sensitivity of the noise-operated squelch to match the new filter. With the addition of the kit, the kit because Regency uses the same i-f board in most of their units; all the holes and spaces for the additional parts and the holes for the new narrowband filter are already there.

HR-2 modifications

If you are one of the many people who own the original HR-2 fm trans-

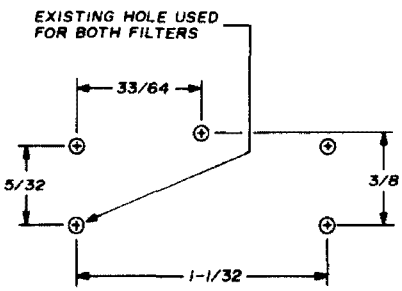


fig. 2. Drilling template for the new narrowband ceramic filter (foil side of board).

ceiver, don't despair. With a little change in the circuit board it is possible to add the modification kit to the earlier models. When the job is finished the i-f board will be electrically identical to the HR-2A.

Circuitwise, the HR-2 and HR-2A are nearly the same in the modification area, but a different circuit board and different parts numbers complicate the instructions supplied with the Regency MA-46 modification kit. The instructions furnished with the kit should be followed, except as noted here. The parts layout on the reverse side of the instruction sheet should *not* be used. Instead, use the information in **fig. 3.**

changes to MA-46 instruction sheet

1a. Remove the old ceramic filter, CF-1

b. Perform the modifications shown in **fig. 1** to the i-f circuit board (301-528-B) and drill the holes for the new ceramic filter using the dimensions given in **fig. 2,** using the existing hole indicated to locate the new holes. Earlier models may have two resistors (R133 and R134) soldered to the foil side of the circuit board as shown in **fig. 4.** If R133 and R134 (both 6.8k) are present, remove them as they are no longer required.

c. Mount the new ceramic filter.

2. Replace the following capacitors with the values indicated.

a. Replace C108 with a 390-pF capacitor.

b. Replace C109 with a 270-pF capacitor.

3. Add C110, a 250-pF capacitor. In early models this capacitor may already be installed, but to assure the correct value, replace any existing C110 with the capacitor furnished in the MA-46 kit.

4. Replace the following resistors with the values indicated

a. Replace R111 with 5.6k resistor.

b. Replace R112 with 2.2k resistor.

c. Replace R137 with a 100 ohm, ¼-watt resistor (not furnished). R137 may be missing on early models. If it is missing it must be added. R137 is located just forward of L103 and installed vertically, as shown to the right (electrically, R137 is connected between C120 and the emitter of Q-102).

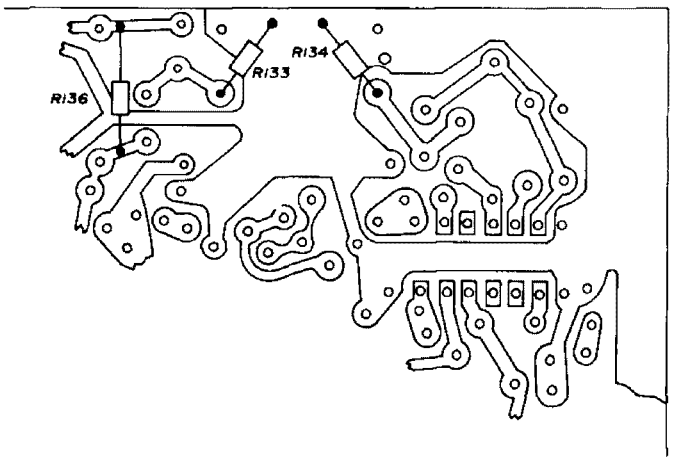


fig. 4. Foil side of the HR-2 receiver i-f circuit board showing the placement of R133, R134 and R136. (R133 and R134 are only present in early models of this fm transceiver.)

d. In early versions of the HR-2, R136 was omitted. In later models it was located on the foil side of the circuit board as shown in **fig. 4.** R136 is a 22k, ¼-watt resistor. If R136 is missing it should be added to the foil side of the circuit board as shown in **fig. 4.** Electrically, R136 is in parallel with L101.

5. Follow the instructions furnished with the MA-46 kit from **step 5** thru to the end of the instruction sheet

With the addition of the MA-46 modifications described here the performance of the Regency HR-2 family of fm transceivers is as good as the latest fm equipment. Furthermore, it can be obtained without the expense of a new rig. The MA-46 modification kit is available from Regency for $22.50, not a bad price when you consider it's almost like getting a brand new receiver, free of that adjacent-channel interference that used to be so annoying.

ham radio

simple high-gain wire antenna for high-frequencies

Design and layout of a four-element, double-extended Zepp that provides up to 7-dB gain on 15 meters

Alvan L. Mitchell, W6QVI, 24765 Calaroga Avenue, Hayward, California 94545

There's an old saying that you can't get something for nothing, especially when you're working with antennas, but you *can* make one wire antenna, the length of a 75-meter dipole, work like a bomb on 75 and deliver 7-dB broadside gain on 15! This is only one-half dB less than a three-element beam on this band. I call the antenna the FEDEZ – Four-Element Double-Extended Zepp.

Many amateurs have used the extended double Zepp which gives 3-dB gain at its design frequency. However, with the addition of phasing stubs and two more elements you can obtain up to 4-dB more gain. All it takes is a little arithmetic which, in my case, was supplied by W6DMY. The basic design was taken from the 1943 edition of the ARRL *Antenna Handbook*. The dimensions for any frequency are given in electrical degrees in fig. 1 (remember that 180° = 1/2 wavelength).

Since most of my on-the-air activities are confined to nets on 75 and 40 meters, with hamming just for fun on 15, the four-element double-extended Zepp I use has a 21.3-MHz center frequency (see fig. 2).

Although the two 7.68-foot phasing stubs can hang straight down from the antenna as shown in fig. 2, I use lumped constants for the two outer stubs as shown in fig. 3. Part of the 450-ohm open-wire feedline is used as the center phasing stub. Each of the lumped-constant stubs I use consist of an 11-turn coil, 2-inches in diameter, 2-3/4 inches long, wound with number-12 wire. Each end of the phasing coil is supported by the strain insulator as shown in fig. 3.

With this antenna I have yet to receive less than an S9 report on the SARO Bourbon net that meets every night on 75 meters, especially from San Diego and Medford, Oregon. On 15 meters I have received numerous S9 reports from the East coast as well as from Japan. WØQWH in Stanley, Kansas, who has given me signal checks on 47 different antennas over the past year, gave me his

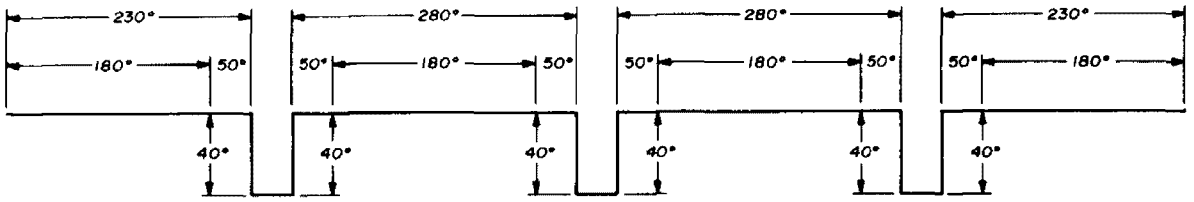


fig. 1. Basic design of the four-element, double-extended Zepp antenna. All dimensions are given in electrical degrees (180° = 1/2 wavelength).

best report, although it wasn't S9 — he apparently has a very stingy S-meter!

The dimensions of my urban lot require that I use this antenna in the inverted-vee configuration. This detracts from the gain somewhat because the wide spacing between the centers of the elements determines gain, and the drooping legs reduce this distance slightly. However, since I feed the antenna with 450-ohm open-wire ladder line to the Ultimate Transmatch,[1] I think I've at last found the ultimate antenna to go with my ultimate transmatch. I don't think you can beat it for city-sized lots.

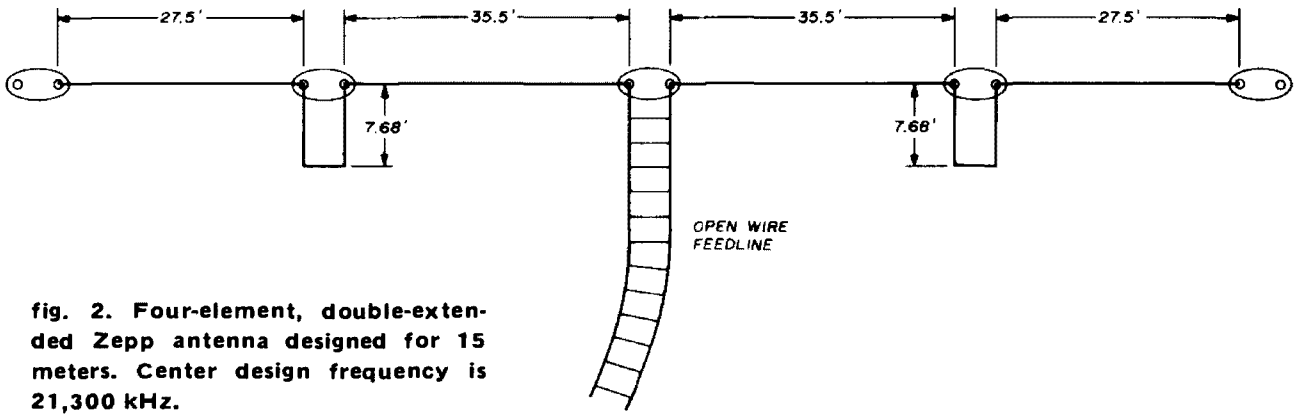


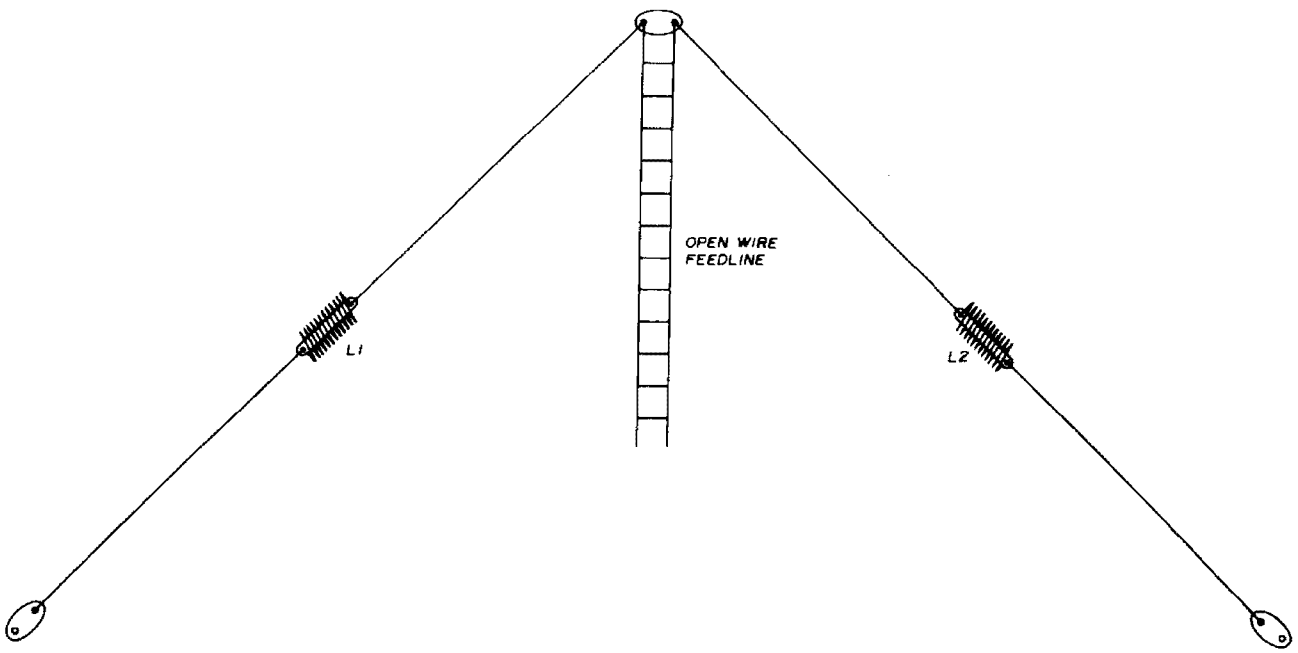
fig. 2. Four-element, double-extended Zepp antenna designed for 15 meters. Center design frequency is 21,300 kHz.

reference

1. Lewis G. McCoy, W1ICP, "The Ultimate Transmatch," *QST*, July, 1970, page 24.

ham radio

fig. 3. The two outer phasing lines can be hung down from the antenna as shown in fig. 2, or phasing inductances may be used as shown here. L1 and L2 are each 11 turns no. 12, 2" diameter, 2-3/4" long. Antenna may be used in the inverted-vee configuration if space is limited.



feedpoint impedance, characteristics of practical antennas

A discussion of antenna feedpoint impedance, and the effects of the resistance and reactance components in practical antennas

Carl C. Drumeller, W5JJ, 5824 N.W. 58 Street, Warr Acres, Oklahoma 73122

The feedpoint impedance, the radiation resistance and dissipative resistance, and the reactance of a common dipole antenna are matters that need clear understanding if you are to inquire deeply into the functioning of that indispensable component of a radio station: the antenna. The purpose of this article is to define and to describe the Z, R and X of antennas, not in a highly technical manner but simply and with only enough detail to distinguish one from another and to show the role each plays.

First, let's consider a center-fed dipole antenna, one a half-wave long (electrically) at the operating frequency, and one out in the clear far enough to have a very minimum modification of its normal characteristics by the influence of its environment. Textbooks tell us that such an antenna will have a feedpoint impedance (Z_f) of 73 ohms, and that this impedance will be purely resistive (no reactance). In the real world, such an antenna seldom is found!

Let's deal first with the ideal dipole, then with the real. In the ideal dipole, Z_f, the feedpoint impedance, will equal R, the composite of the radiation resistance and each of all of the dissipative resistances. These dissipative resistances include the ohmic resistance of the antenna, insulation losses, dielectric losses and absorption losses. These are easy to visualize. You know that the antenna wire has

resistance, even though it's made of highly-conductive copper. You know that no insulator is perfect; so even the best has some loss. You know that somewhere within the near-field of the antenna there must be an insulating object that introduces dielectric losses, however small. And you know that somewhere within the near-field there must be some material that will absorb radio waves.

radiation resistance

Radiation resistance, though, is a different matter! In the first place, it's not a true resistance. It acts like a resistance in some ways, but not in every manner. For instance, a real resistance, when radio-frequency current flows through it, converts the electrical energy into heat, another form of energy. Radiation resistance doesn't do this.

What, then, does it do? Nothing! It's just a term which describes an attribute of an antenna, an attribute which bears a superficial resemblance to a real resistor.

The need for such a term comes about from the fact that all of the rf power that flows into an antenna doesn't get converted into heat. Some (and, we hope, a greater part) of the rf power is radiated out into space. It's convenient to speak of an antenna's characteristics as if all of the rf energy fed into it were dissipated just like that portion which produces heat. To make this fiction plausible, we assign an imaginary resistor to the antenna and call it "the radiation resistance."

When we put a known amount of rf power into the antenna, defining it as $W=I^2R$, and having a known amount of current, we have a *large* enough value of resistance to make the formula valid. We've taken care of not only the amount of power that was dissipated in the various real resistances *and* equivalent resistances but also the amount of power radiated into space; the latter being equal to what a real resistance of a value the same as the *radiation* resistance would have dissipated in the form of heat.

Let's run that through again. Taking a purely imaginary situation, let's conjure up an antenna that has only real resistance, a real resistance of one ohm, and feed one ampere of rf current into it. According to the formula, only one watt of power is going into that antenna, and all of it is being converted into heat with none of it being radiated.

Now, conjure up another antenna with one ohm of real resistance and 49 ohms of radiation resistance. With the same one-ampere of rf current going into it, the formula tells us that 50 watts of rf power is going into the antenna — one watt is squandered as heat and 49 watts are radiated. Quite an improvement!

This brings us to a cardinal rule: Make the ratio of radiation resistance to dissipative resistance as high as you can. This is not too difficult to do with a half-wave or even a quarter-wave antenna, but when you attempt it with a really small antenna, say a tenth-wave, you run into a real problem. That's why the engineers who design 80-meter mobile antennas work up such a sweat over their drawing boards.

So much for radiation resistance. Just remember that it's an imaginary resistance that accounts for the power being radiated by the antenna.

antenna reactance

Now for the reactance. Remember, we started out with an ideal antenna, one that was resonant and therefore resistive. It might be resonant on, say, 7,257,376 Hz, but when you breathe on your transmitter and it drifts to 7,257,377 Hz, the antenna departs ever so little from resonance. As it departs from resonance, it loses that purely resistive status. If the frequency goes higher, a bit of inductive reactance is introduced; if it goes lower, the introduced reactance is capacitive.

Just how much the antenna departs *from resistive to* resistive-plus-reactive status, or, rather, the rate at which it departs for a given change of frequency, depends upon several factors. For the simple dipole we're considering, the chief of these factors is the antenna's diameter to length ratio. The larger the diameter of the radiator for a given length, the less

reactance introduced for a given change of frequency. The almost-obsolete cage antenna merits much consideration, for it gives a very favorable ratio of diameter to length.

When reactance is present the feedpoint impedance, Z_f, no longer equals R. It is given by

$$Z_f = V\sqrt{R^2 + X^2}$$

with the R still the grand total of all the resistances (radiation, ohmic, etc.) and X either inductive ($+X_L$) or capacitive ($-X_C$), as the case may be. In either case when X is squared it's a positive value, so forget the sign.

There's one thing you musn't forget, though. That's the matter of reactance not being able to absorb power. Ponder this, for it's quite important! Think of what it involves. The feedpoint impedance may go high and you feel that the dissipative resistance of your antenna is low. You rejoice, believing you're radiating more power, a valid assumption only if it were the radiation resistance that was going up. You can't make a purely-reactive termination accept power. One that's partly-reactive and partly-resistive, yes. One that's purely reactive, no.

Don't jump to the conclusion that reactance in an antenna is an evil thing. In certain antenna designs it plays a vital role, but this is not an article on antenna design. If you want to look into that subject, get a reliable textbook, preferably one written by Kraus, LaPorte, or some other recognized authority on the subject. There's a wide difference between the simple dipole we're discussing and a complex antenna. For this article we'll stick to the dipole!

If your dipole is reactive to a degree, as are the vast majority of such antennas, don't worry about it. If it does give you concern, remember that the reactance can be cancelled out by the introduction of an equal and opposite reactance. For example, if the antenna exhibits 10 ohms capacitive reactance, this can be negated by introducing 10 ohms of inductive reactance. This conjugate reactance can be placed at the feedpoint of the antenna or at any point between that feedpoint and the active device in your transmitter. Its position doesn't matter so long as it reflects that conjugate reactance into the antenna. Keep in mind that the resistive component of the antenna's impedance, which will not be affected by these manipulations to cancel reactance, is going to accept the rf power.

resistance transformation

The resistive component can be transformed by many and various means to any convenient numerical value that you might elect to stipulate. Again, this can be done at the feedpoint of the antenna or at any place between that feedpoint and the active device in your transmitter.

In each instance, there is some slight advantage in having the transformation take place at the antenna's feedpoint. With some transmitters, ones poorly designed or manufactured to meet a price and not to provide quality, it is imperative that the transformation take place between the antenna's feedpoint and the transmitter's antenna terminal. This, though, is strictly a transmitter deficiency.

summary

To sum up, the feedpoint impedance of an antenna is a complex quantity, consistituted by both resistive and reactive components. The resistance component may be made up of many constituents. Of these, one, the radiation resistance, is not a true resistance but an imaginary one invented to account for the rf energy radiated by the antenna. The several other constituents of antenna resistance are all dissipative in nature and should be held to a minimum in design. Radiation resistance should be high as compared to the total of the other resistances. Some element of reactance is present in most antennas, but this is not a significant deficiency and may even be used to advantage in some designs.

ham radio

improved logic test probe

This improved logic test probe checks binary levels as well as pulse coincidence

R.H. Fransen, VE6RF, 227 Cottonwood Avenue, Sherwood Park, Alberta, Canada

Since I have always been interested in test equipment, the TTL logic probe with a built-in memory described in a recent issue of *ham radio*[1] proved very interesting. I made some changes to the basic circuit so that it can take the place, in many instances, of an item we would all like to own but can't afford, a dual-trace oscilloscope. The design uses three ICs, some additional switches and more hardware.

Since I wasn't able to obtain some of the parts used in the original logic probe, like good, bright LEDs, some circuit changes were made as needed. The completed unit may look a little clumsy in its mechanical design because I used what was available, but the probe does the job it's supposed to do, and that's what counts. If you have access to better materials you can dress it up any way you like.

the circuit

The logic probe circuit, **fig. 1**, has two inputs, *main* and *auxiliary*. In the off position of the *off-aux* switch the unit operates as in the original design. However, in this circuit you can switch the memory off with the *off-mem* switch so you don't have to keep pushing the button when using the probe as a binary level indicator.

In the *aux* position of the *off-aux* switch two inputs are needed at the same time. The level of the pulse into the *aux* jack is selected by the *aux* + or - switch (see **fig. 2**). To check the coincidence of pulses, just connect a patch cord from the *aux* jack to the second point on the logic circuit you are checking, and the probe will indicate it.

The parallel RC circuit in series with the *aux* input is to protect the probe against a direct short to common in case the *aux* input is connected and the *off-aux* switch is in the *off* position. The

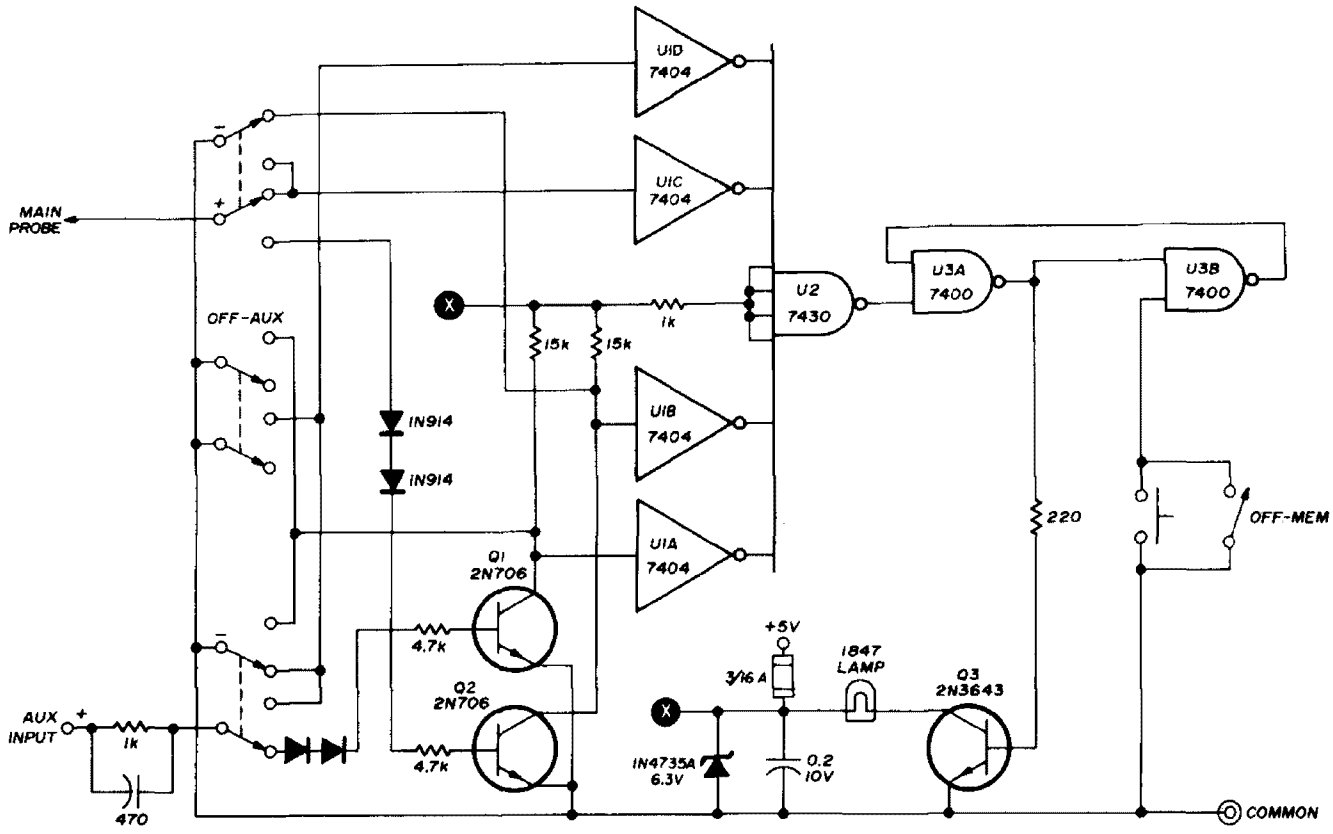


fig. 1. Circuit for the improved logic test probe. All signal diodes are 1N914, all resistors are ½ watt.

470-pF capacitor prevents too much pulse slow down. The 1N914 diodes serve to bring up the high trigger threshold voltage to prevent noise triggering.

I was unable to obtain a decent, bright LED, so I used a long-life number-47 bulb with a switching transistor. At the voltage used, the bulb should last forever, and it's still bright enough to be seen, even in bright sunlight.

The *common* of the TTL circuit is connected to a pin jack for those cases where the circuit under test cannot handle the probe current requirements. With the 1847 bulb the probe needs a total of about 160 mA; changing to a LED would cut probe current to 60 mA.

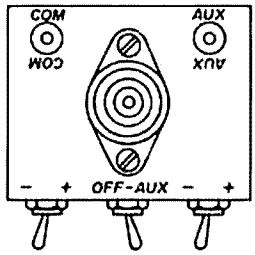
construction

The entire logic probe circuit is built into a 3¼x2¼x1½-inch aluminum box similar to a Minibox, and fed by a small audio-type coaxial cable with alligator clips on the opposite end. The main probe is mounted on a small ceramic stand-off insulator. The 0.2-μF capacitor is mounted on a small section of Vero board right where the feed cable is connected. The Vero board with all the ICs and other components is mounted on the inside of the aluminum box. All the ICs are mounted in sockets. (It is especially important to mount the 7404 IC in a socket since it serves as a cheap fuse and will burn out if too high voltages are applied to any of the probe inputs.)

operation

In use, the metal case of the logic probe is left floating. Supply current with the lamp off is 26 mA; with the lamp on, current drain is 160 mA. *Main* trigger threshold voltages are +1.5 volts (high) and +1.3 volts (low). *Auxiliary* trigger threshold voltages are 1.5 volts (high) and 0.7 volt (low). The *aux* input can be used by itself if the *main* input is switched to minus (−) and connected to *common.* This may be useful at times since the low level of the *aux* input is half as low as the low level on the *main* input.

If a separate power supply is used for the logic probe only the *common* of the probe must be connected to the negative line of the TTL circuit under test. It should also be kept in mind that when



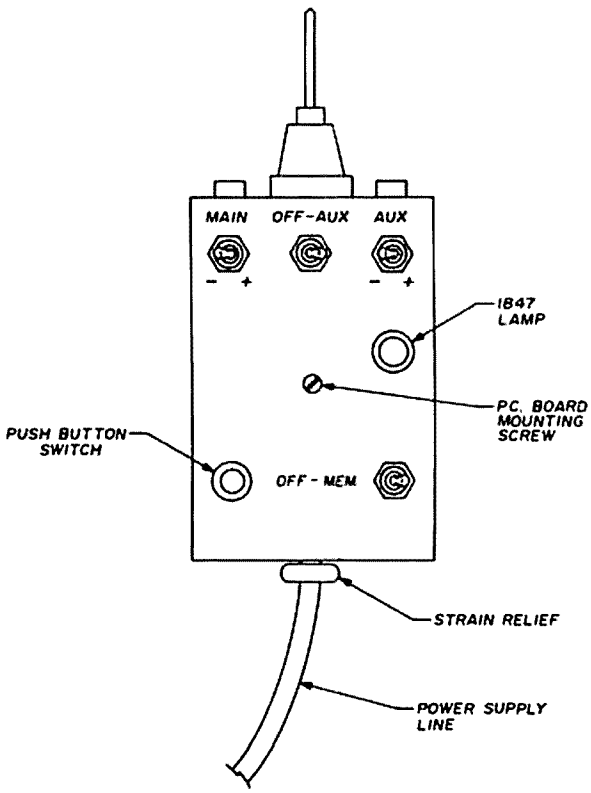


fig. 2. Construction of the improved logic test probe. Unit is housed in a small aluminum box; power supply is external.

checking TTL pulse trains with the memory switched off, and the indicator does not dim, it is probably because the duty cycle of the pulse is not 50%. Switching the polarity with the plus/minus switch may show more dimming than usual as with a 50% duty cycle. Experience will quickly show what to expect.

If a separate power supply is needed, a transformer, some diodes, a filter capacitor and one of the new 5-volt IC regulators will do the trick.

reference

1. W. Rossman, "Logic Test Probe," *ham radio*, February, 1973, page 56.

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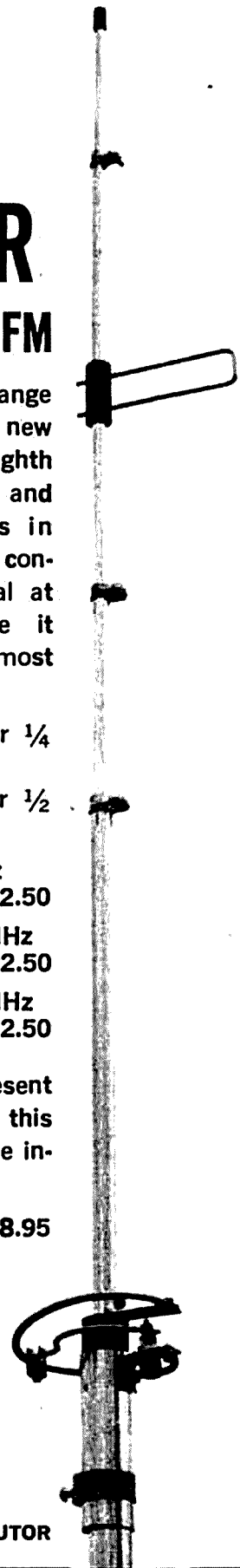
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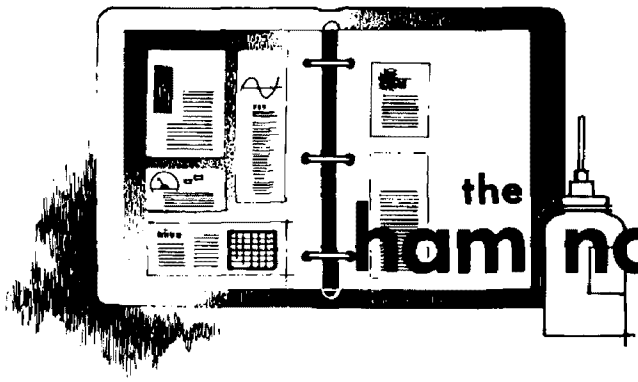
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the ham notebook

TTL clock oscillator

In the circuit shown, two IC one-shot multivibrators are cross-coupled to make an oscillator suitable for driving other TTL ICs for various logic applications. The outputs are somewhat more TTL compatible than those obtained using transistor or unijunction circuitry.

In addition, this circuit is well suited to applications where the clock must be started and stopped at suitable intervals. In fact, it is necessary to have at least one positive-going transition on the enable input to start the clock after power is applied. The circuit by itself will not free run simply by applying a logic one level to the enable input. Note that both one-shots must time out after the enable goes low before the clock comes to rest.

The output of the first one-shot produces a pulse immediately after the enable input goes high, while the second one-shot waits until the end of the first cycle before it produces a pulse. The duty cycle of the output waveform can be adjusted as required by making both timing resistors variable. These also set the frequency of the oscillator.

With the IC one-shots, both the normal and inverted outputs of the clock are available at the "Q" and "not Q" terminals. If RCt_2 is made a very short duration pulse and RCt_1 is made adjustable over a wide range, a variable frequency pulse train of thin widths is produced. Making the two time constants equal produces a square wave output.

Cal Sondgeroth, W9ZTK

yaesu sideband switching

For owners of the Yaesu Ft-101 who miss the convenience of switching sidebands without retuning, here's a simple modification which can be made without affecting any other function of this fine equipment.

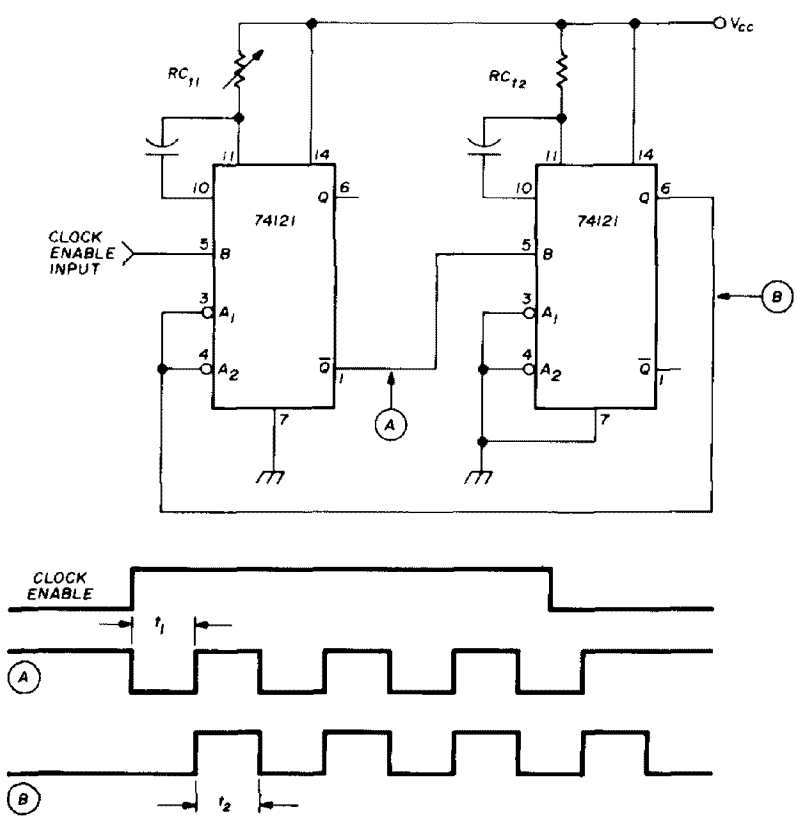


fig. 1. Simple clock-oscillator circuit using two TTL 74121 monostable multivibrator ICs.

By taking advantage of the clarifier circuitry and adding a potentiometer between two of the circuit-board receptables, MJ6-11 and MJ5-2, an adjustment can be made which puts the vfo frequency in the right spot when switching to upper sideband, tune or CW. A small piece of perfboard, a 2500-ohm PC-mounting pot (39 cents from Radio

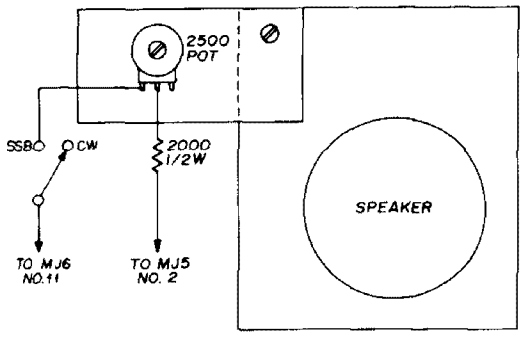


fig. 2. The 2500-ohm pot and 2k resistor zero the carrier frequency when shifting sidebands with the Yaesu Ft-101. Location is not critical as all circuitry is dc. Be sure, however, that the screw-adjust pot can be reached with the bottom shelf cover in place.

Shack) and a 2000-ohm, ½-watt resistor wired in series are the only things needed.

To align the circuit after it is installed, tune in a 3800-kHz lower-sideband signal and zero-beat the calibrator signal to it. Now, switch to upper sideband and adjust the 2500-ohm pot for zero beat. That completes the alignment. The setting at the center of the vfo range holds within a few Hz throughout the tuning range of the vfo, and is the same for USB, tune or CW. It is a pleasure when switching sidebands or going from CW to ssb not to have to recalibrate the dial.

Ernie Schultz, W2MUU

nuvistor heatsinks

Transistor-type heatsinks make excellent heat-dissipating radiators for nuvistor-type vacuum tubes. If possible, choose a high emissivity black anodized heatsink and use thermal compound between the metal tube and the heatsink to maximize heat transfer.

Richard Mollentine, WAØKKC

exploding diodes

If you have done much experimenting with the very popular glass encapsulated diodes you will know that they tend to explode rather violently when subjected to a severe overload. Since most amateurs and experimenters don't wear safety glasses, this could be a dangerous situation. When these glass diodes explode they blow very small fragments of broken glass over a considerable area with enough force to cause serious eye injury. To prevent this from happening when experimenting and building projects, take a small piece of Scotch tape and wrap it tightly around each glass diode before installation. If you accidentally short something the tape will contain the force of the explosion and prevent the glass from blowing all over the room, possibly saving someone's eye.

Pete Walton, VE3FEZ

Heathkit HW-16 problems

While repairing a Heathkit HW-16 Novice transceiver, I found the answer to several problems which may have bothered others. Keying characteristics were harsh, with pronounced clicks. A capacitor up to half a microfarad across the key, in parallel with C92, helped greatly. A .01-μF ceramic capacitor across R14 also helped to keep down QRM in the novice band.

The sidetone oscillator, a neon bulb, lit, but refused to oscillate. A larger resistor in place of R64 took care of this. I changed the original value of 1.5 megohms to 3 megohms. Varying this resistor also changes the tone.

The meter read half-way up scale with no current through it. Investigation showed the metal band around the plastic meter case was magnetized. A careful application of a magnetized screwdriver reversed this condition, and after several trials, the pointer rested near zero, where it should.

Eugene A. Hubbell, W7DI

short circuits

HW16 modification

In the March, 1973, issue of *ham radio*, a 0.001-μF blocking capacitor should be placed in series with the shielded lead connected from the grid of V7 to the grid of V2A. A number of readers have complained of insufficient power on 15 meters, but WB6MZN, the author, indicates that his plate power meter reads 160 mA on 40 meters and 180 mA on 15. He points out, however, that in the original configuration the HW16 tends to oscillate and power decreases on 15 meters. He cured this by carefully tuning all the tank circuits especially for 15-meter operation, including L8 and C21, the neutralizing capacitor.

ac power supply for fm equipment

In the ac power supply on page 28 of the June, 1973, issue the regulator transistor may oscillate under certain load conditions. This oscillation can be suppressed by installing a 0.47-μF bypass capacitor from the output to ground. When paralleling power transistors for greater current capacity, be sure to include 0.1-ohm, 2-watt balancing resistors in series with the emitters of *each* power transistor.

1296-MHz quad yagi

In the May, 1973, issue the driven element for the 1296-MHz quad Yagi should be made from 1/32-inch-thick flat brass. The reflector and directors are made from flat aluminum stock, 0.050-inch thick, *not* rod as stated in the article.

phase II receiver

There were several circuit errors in the Phase II Receiver published in the August, 1973, issue of *ham radio*. In **fig. 3** R35 should have a value of 100k ohms and C33 is not used on the PC board at all. The jumper just below U6 in **fig. 4** should be connected to the circuit pad at the lower right hand corner of U6 (goes to pin 2 not to the pad on pin 15). Rf choke L1 is approximately 6 μH and may be wound on an Amidon T37-2 core.

The author reports that the mosfet, Q4, suffers from parasitics and is touchy to agc. The whole stage may be replaced with an emitter follower (2N3707) with a 680-ohm emitter resistor and a 470k base-bias resistor. The 100k agc control, R37, may be replaced with a 100k fixed resistor. True rf gain control can be obtained by replacing CR1 with a 1000-ohm pot. Reduced gain results in better cross-mod performance. The author has inserted two 1N914 diodes in the agc line running to Q1 to improve strong-signal performance. With only Q1 controlled, agc range is about 40 dB and much smoother. This range depends upon the setting of the new 1k rf gain control.

The dc offset to U3 (MC1741CG) may need to be adjusted if the quiescent voltage at pin 6 is not near 5 volts.

micropower receiver

In the schematic for the micropower communications receiver in the June, 1973, issue of *ham radio*, a 220k base bias resistor should be connected from the base of the 2N1307 transistor to the +6 volt supply line.

motorola test set

In the Motorola test set article in the November, 1973, issue of *ham radio*, it should be noted that in late model Motrac, Motrans, Mocom and Micor radios the first i-f has been changed from 12 MHz to 8 MHz. When aligning the first i-f it must be determined which frequency is involved. If the first i-f adjustments are tuning capacitors, the i-f is a 12-MHz unit. If the adjustments are slug-tuned coils, the i-f is at 8 MHz.

logic test probe

In the circuit for the logic test probe featured in the *ham notebook* section in the February, 1973, issue, no power connections were shown for the IC. Connect +5 volts to pin 14 and ground pin 7 of the IC.

antenna and control-link calculations

The appendix for W7PUG's "Antenna and Control-Link Calculations" article in the November, 1973, issue of *ham radio* was inadvertently not included with the article. The *Tymshare Superfortran* program for antenna pattern calculations is shown in table 1, below. Examples of computer printouts for the two types of antennas discussed in the article are shown in tables 2 and 3.

table 1. Tymeshare Superfortran computer program for calculating antenna patterns. Sample computer printouts for a J-pole and Stationmaster antenna are shown in tables 2 and 3, respectively.

```
100           DIMENSION EE(90),EH(90)
101           PI=3.14159265
102           RD=PI/180.
103           ACCEPT'NUMBER OF COLINEAR ELEMENTS: ',N
104           ACCEPT'ELEMENT SPACING IN WAVELENGTHS: ',DW
105           ACCEPT'OFFSET FROM REFLECTING MOUNT, WAVELENGTHS: ',OW
106           SEH=0.
107           DO 10 I=1,36
108           EH(I)=COS(2.*PI*OW*(COS(5.*RD*I)-1.))
109           SEH=SEH+RD*EH(I)*EH(I)
110     10    CONTINUE
111           SEE=0.
112           DO 20 I=1,90
113           PSI=PI*DW*SIN(RD*I)
114           EE(I)=COS(RD*I)*SIN(N*PSI)/(N*SIN(PSI))
115           SEE=SEE+RD*COS(RD*I)*EE(I)*EE(I)
116     20    CONTINUE
117           GH=10.*LOG10(.2*PI/SEH)
118           GE=-10*LOG10(1.64*SEE)
119           GMAX=GH+GE
120           WRITE(1,100)GH,GE,GMAX
121     100   FORMAT(/'H-PLANE GAIN',F6.1,' DB, E-PLANE GAIN',F6.1,
              ' DB'/'MAXIMUM MAIN LOBE GAIN IS',F6.1,' DB.'/)
122           IF(OW.EW.0.)GOTO 40
123           DISPLAY'   AZ.    GAIN   RELATIVE'
124           DISPLAY'   DEG.    DB    VOLTAGE'
125           WRITE(1,101)GMAX
126     101   FORMAT('     0',F7.1,'     1.00')
127           DO 30 I=2,36,2
128           J=5*I
129           DB=GMAX+10.*LOG10(EH(I)*EH(I))
130           WRITE(1,102)J,DB,EH(I)
131     102   FORMAT(I5,F7.1,F8.2)
132     30    CONTINUE
133     40    DISPLAY''
134           DISPLAY'   ELEV   GAIN   RELATIVE'
135           DISPLAY'   DEG.    DB    VOLTAGE'
136           WRITE(1,101)GMAX
137           DO 50 I=1,90
138           DB=GMAX+10.*LOG10(EE(I)*EE(I))
139           WRITE(1,102)I,DB,EE(I)
140           IF(I.GE.20)I=I+4
141     50    CONTINUE
142           STOP
143           END
```

table 2. Computer-generated antenna pattern information for a 4-element J-pole antenna.

```
NUMBER OF COLINEAR ELEMENTS: 4
ELEMENT SPACING IN WAVELENGTHS: 1
OFFSET FROM REFLECTING MOUNT, WAVELENGTHS: 0.1

H-PLANE GAIN    2.3 DB, E-PLANE GAIN    6.5 DB
MAXIMUM MAIN LOBE GAIN IS    8.9 DB.


ELEV   GAIN   RELATIVE        AZ.   GAIN   RELATIVE
DEG.    DB    VOLTAGE        DEG.   DB    VOLTAGE
  0     8.9    1.00            0    8.9    1.00
  1     8.8     .99           10    8.9    1.00
  2     8.6     .97           20    8.9    1.00
  3     8.3     .93           30    8.8    1.00
  4     7.8     .88           40    8.8     .99
  5     7.1     .82           50    8.6     .97
  6     6.3     .75           60    8.4     .95
  7     5.3     .66           70    8.1     .92
  8     4.1     .58           80    7.6     .87
  9     2.5     .48           90    7.0     .81
 10      .7     .39          100    6.2     .74
 11    -1.8     .29          110    5.3     .67
 12    -5.0     .20          120    4.2     .59
 13    -9.8     .12          130    3.1     .51
 14   -20.1     .04          140    1.8     .44
 15   -19.8    -.04          150     .6     .39
 16   -11.1    -.10          160    -.4     .34
 17    -7.5    -.15          170   -1.1     .32
 18    -5.4    -.19          180   -1.3     .31
 19    -4.1    -.23
 20    -3.4    -.24
 25    -5.4    -.19
 30  -161.6    -.00
 35    -6.6     .17
 40    -4.8     .21
 45   -10.0     .11
 50   -17.5    -.05
 55    -5.0    -.20
 60    -1.5    -.30
 65     -.6    -.34
 70    -1.3    -.31
 75    -3.1    -.85
 80    -6.4    -.17
 85   -12.3    -.09
 90  -166.0    -.00
```

table 3. Computer-generated antenna pattern information for a type-2 antenna (Communications Products Stationmaster).

```
NUMBER OF COLINEAR ELEMENTS: 9
ELEMENT SPACING IN WAVELENGTHS: 0.3
OFFSET FROM REFLECTING MOUNT, WAVELENGTHS: 0

H-PLANE GAIN     .0 DB, E-PLANE GAIN    5.7 DB
MAXIMUM MAIN LOBE GAIN IS    5.7 DB.

          ELEV   GAIN   RELATIVE
          DEG.    DB    VOLTAGE
            0     5.7    1.00
            1     5.7    1.00
            2     5.6     .99
            3     5.4     .97
            4     5.2     .94
            5     4.9     .91
            6     4.5     .87
            7     4.1     .83
            8     3.5     .78
            9     2.9     .72
           10     2.2     .67
           11     1.4     .61
           12      .5     .55
           13     -.6     .49
           14    -1.8     .42
           15    -3.2     .36
           16    -4.8     .30
           17    -6.7     .24
           18    -9.0     .18
           19   -12.1     .13
           20   -16.4     .08
           25   -13.4    -.11
           30    -8.8    -.19
           35    -9.4    -.17
           40   -13.4    -.11
           45   -23.2    -.04
           50   -27.1     .02
           55   -19.3     .06
           60   -17.8     .07
           65   -18.5     .06
           70   -20.5     .05
           75   -23.6     .03
           80   -27.8     .02
           85   -34.4     .01
           90  -188.2     .00
```

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